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## EDITORIAL

### Amplitude and Frequency Modulation

IN this number we publish an article sent to us from the U.S.A. by M. G. Nicholson in which the relative merits of the two methods of modulation are compared. This article should serve as a warning to those who are under the impression that frequency modulation is greatly superior to amplitude modulation in most, if not in all, circumstances. One is apt to gain such an impression from the campaign that is being waged on behalf of frequency modulation, especially in America. Mr. Nicholson says that much confusion is caused by comparison between the two under entirely differents sets of conditions. When the conditions are similar he maintains that there is surprisingly little to choose between them. It is obviously useless to compare results obtained with amplitude modulation at medium frequencies with those obtained with frequency modulation at very high frequencies; the comparisons must be made under similar conditions. The article is based on both theoretical and practical results.

In comparing the fidelity of reproduction, better results are naturally obtained with f.m. at a frequency of 50 Mc/s and a channel width of 200 kc/s than with a.m. at 1 Mc/s and a channel width of only 10 kc/s, but this must not be put to the credit of the method of modulation but to that of the high frequency which makes the greater channel width possible.

The author discusses the effects of all the various forms of interference and, as can be seen at once from his conclusions, finds that only where fluctuation noise is the limiting factor is frequency modulation superior to amplitude modulation; in all other cases he maintains that the latter is the better of the two. This is very important in view of the frequent complaints of the interference caused by motor-car ignition systems. The author goes very fully into such impulse interference and comes to the conclusion that amplitude modulation is superior to frequency modulation in discriminating against it.

Considerable attention is devoted to the subjects of pre-emphasis and de-emphasis which are essential in f.m. in order to reduce the excessive noise at high audio frequencies, and which consequently give a greater improvement in f.m. than in a.m.

Even if some of the results are queried, it is obvious that any organisation proposing to install and test frequency modulation must, if the results are to be of any value, also install and test amplitude modulation of the same, or a similar, transmitting station under identical conditions. If Mr. Nicholson's arguments and experimental results are even approximately correct, there should be surprisingly little difference in the final performance.

G.W.O.H.

# Vicalloy—A Workable Alloy for Permanent Magnets

A MONOGRAPH by E. A. Nesbitt with this title has recently been issued by the Bell Telephone System. Phenomenal advances have been made in the last sixteen years in materials suitable for permanent magnets. In the Editorial of July 1942 we discussed some of these advances and pointed out that the mechanical properties of the new alloys of the Alnico or Ticonal type had revolutionized production methods, since they cannot be turned, milled, drilled or tapped, and the magnets have consequently to be cast and ground. This is due to the excessive brittleness of these alloys. In the new alloy called Vicalloy (v=vanadium, i=iron, c=cobalt) this disadvantage is eliminated to a large extent and the material can be rolled and drawn. It has been used commercially as a tape 0.002 in by 0.05 in for the recording of speech in the Western Electric Mirrorphone. The range of percentage composition is given as—iron 30 to 52, cobalt 36 to 62, vanadium 4 to 16. In many ways the properties of the alloy are quite contrary to what one would expect from experience with the old-fashioned carbon steels. After quenching in oil from 1,200° C the residual $B_r$ was 4,950 and the coercive force $H_c$, 51; after aging—one can scarcely call it annealing—for 8 hours at 600° C $B_r$ had increased to 9,000 and $H_c$ to 300; in fact this aging process is necessary to give it its permanent magnet properties. Another strange result is that, although Vicalloy with 14 per cent vanadium is almost non-magnetic before it is cold-worked, by rolling and drawing it cold until its cross-section is reduced to 5 per cent of its original value and then aging at 600° C one obtains a material with the very high $(BH)$ maximum value of $2.8 \times 10^6$ in the longitudinal direction. In the transverse direction the value of $(BH)_m$ would be only a fraction of this.

Although the magnetic properties of Vicalloy are inferior to those obtainable with Ticonal, its mechanical properties, which allow it to be machined, punched, drilled, and tapped, will in many applications more than compensate for its magnetic inferiority.

G. W. O. H.

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# TRANSMISSION IN WAVEGUIDES*

## Cross-section Partly of Solid Dielectric

*By Alice M. Woodward, M.A.*

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**SUMMARY.**—The theory of transmission of an $H_{01}$ wave† in a rectangular wave-guide containing longitudinal slabs of solid dielectric is developed. Formulae are given for phase constant and attenuation which take into account imperfect conductivity of the guide walls as well as losses in the dielectric. Numerical values are given for polythene as dielectric. At high frequencies most of the energy travelling down the guide is confined to the dielectric slabs. The phase constant and attenuation are then nearly equal to their values for a completely-filled guide.

### Introduction

MANY papers[1, 2, 3, 4, 5] have now been written on the theory of the transmission of electromagnetic waves in waveguides completely-filled with a dielectric material. They show that the phase velocity of the wave depends on its frequency and on the dielectric material used, being greatest for a guide filled with air. It is evident then, that the phase-velocity of a wave of any given frequency can be controlled by varying the dielectric constant of the material in the guide or, as this is not always convenient, by using two dielectrics in varying proportions. Energy considerations prompt the use of air as one of the dielectrics, since the conductivity of a solid dielectric leads to quite considerable losses. Moreover, it has

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† An $H$-wave in a guide is defined as one in which there is no electric vector in the direction of the length of the guide.

been pointed out[5] that the losses in the conducting walls of an air-filled guide are less than those in a guide filled with any other dielectric. Polythene is a convenient material for the other dielectric.

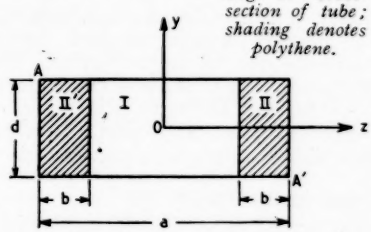


Fig. 1. *Cross-section of tube; shading denotes polythene.*

This paper, which was first prepared as a T.R.E. Report in October 1941, deals with a rectangular waveguide partially filled with polythene, its cross-section being as indicated in Fig. 1. An $H_{01}$-wave is propagated down the guide, its electric vector parallel to Oy. In Sections 2 and 3, the phase-constant and the attenuation due to losses in the polythene are derived on the assumption that the walls of the guide are perfectly conducting, as their imperfect conductivity makes a negligible correction. The attenuation of the wave due to losses in the imperfectly conducting walls is taken account of separately in Section 4.

## 2. Phase-Constant in a Perfectly-Conducting Guide

In Fig. 1, AA′ represents the cross-section of a conducting pipe of great length perpendicular to the paper. Regions II and II′ are filled with a solid dielectric of dielectric constant $\kappa_1$, permeability $\mu_1$ and conductivity $\sigma_1$. Region I is air-filled, with dielectric constant $\kappa_0$, and permeability $\mu_0$.

Suppose an $H_{01}$-wave to be travelling down the guide. The field-intensities of electromagnetic waves in pipes have been derived elsewhere and it is sufficient to state here that, in Region I,

$$\left.\begin{aligned} E_y &= A\cos\{z\sqrt{k_0^2+P^2}\}\,e^{-Px+j\omega t} \\ H_z &= \frac{P_1}{j\omega\mu_0}\times \\ &\quad A\cos\{z\sqrt{k_0^2+P^2}\}\,e^{-Px+j\omega t} \\ H_x &= -\frac{\sqrt{k_0^2+P^2}}{j\omega\mu_0}\times \\ &\quad A\sin\{z\sqrt{k_0^2+P^2}\}\,e^{-Px+j\omega t} \end{aligned}\right\} \quad 2.1$$

where $k_0^2=\omega^2\mu_0\kappa_0$, and $\omega = 2\pi\times$ frequency: $P$, the propagation constant $=\alpha+j\beta$.

Assuming, at first, that the tube is a perfect conductor, the fields in II are, likewise,

$$\left.\begin{aligned} E_y &= B\sin\{(\tfrac{1}{2}a-z)\sqrt{k_1^2+P^2}\}\,e^{-Px+j\omega t} \\ H_z &= \frac{P}{j\omega\mu_1}B\sin\{(\tfrac{1}{2}a-z)\sqrt{k_1^2+P^2}\}\,e^{-Px+j\omega t} \\ H_x &= -\frac{\sqrt{k_1^2+P^2}}{j\omega\mu_1}B\cos\{(\tfrac{1}{2}a-z)\sqrt{k_1^2+P^2}\}\,e^{-Px+j\omega t} \end{aligned}\right\} \quad .. \quad 2.2$$

where $k_1^2=\omega^2\mu_1\kappa_1-j\omega\mu_1\sigma_1$.

The propagation constant, $P$, is obtained from the boundary conditions at the interface of the air and the dielectric where $E_y$ and $H_x$ must be matched. This gives

$$A\cos\{(\tfrac{1}{2}a-b)\sqrt{k_0^2+P^2}\} = B\sin\{b\sqrt{k_1^2+P^2}\}\times \frac{\sqrt{k_0^2+P^2}}{j\omega\mu_0}A\sin\{(\tfrac{1}{2}a-b)\sqrt{k_0^2+P^2}\} = \frac{\sqrt{k_1^2+P^2}}{j\omega\mu_1}B\cos\{b\sqrt{k_1^2+P^2}\},$$

yielding

$$\frac{\sqrt{k_0^2+P^2}}{\mu_0}\tan\{(\tfrac{1}{2}a-b)\sqrt{k_0^2+P^2}\} = \frac{\sqrt{k_1^2+P}}{\mu_1}\operatorname{cotan}\{b\sqrt{k_1^2+P^2}\} \quad .. \quad .. \quad .. \quad 2.3$$

Neglecting $\sigma_1$ initially, so that $\alpha=0$, Equation 2.3, rewritten with $P=j\beta$ is

$$\left.\begin{aligned} \frac{\sqrt{k_0^2-\beta^2}}{\mu_0}\tan\{(\tfrac{1}{2}a-b)\sqrt{k_0^2-\beta^2}\} &= \frac{\sqrt{k_1^2-\beta^2}}{\mu_1}\operatorname{cotan}\{b\sqrt{k_1^2-\beta^2}\},\ \text{for } \beta^2<k_0^2 \\ -\frac{\sqrt{\beta^2-k_0^2}}{\mu_0}\tanh\{(\tfrac{1}{2}a-b)\sqrt{\beta^2-k_0^2}\} &= \frac{\sqrt{k_1^2-\beta^2}}{\mu_1}\operatorname{cotan}\{b\sqrt{k_1^2-\beta^2}\},\ \text{for } k_0^2<\beta^2<k_1^2 \end{aligned}\right\} \quad 2.4$$

Equations 2.4 may be solved graphically for $\beta$, the first branches of the curves giving $\beta$ for the $H_{01}$ wave. Subsequent branches give higher order waves. The product of tube-width, $a$, and cut-off frequency is obtained from Equation 2.4 by putting $\beta = 0$, and is plotted against $b/a$ in Fig. 2.

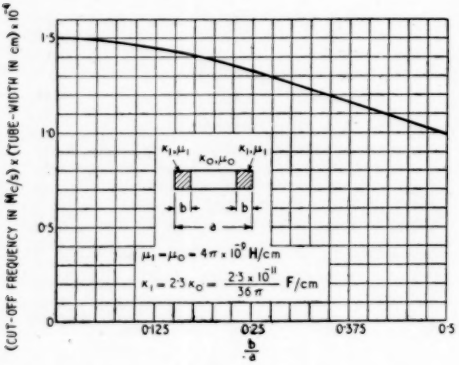


Fig. 2. *Cut-off frequency.*

Fig. 3 gives graphs, calculated from Equation 2.4, of $a\beta$ against the product of $a$ and frequency for

$$\frac{b}{a} = 0, \frac{1}{8}, \frac{1}{4}, \frac{3}{8}, \frac{1}{2}.$$

The solid dielectric used is polythene, for which the following constants are assumed:

$$\mu_1 = \mu_0 = 4\pi \cdot 10^{-9} \text{ henrys/cm}$$
$$\kappa_1 = 2.3\kappa_0 = 2.3 \cdot 10^{-11}/36\pi \text{ farads/cm.}$$

It is interesting to notice that the curves tend to follow that for $\frac{b}{a} = 0$ (the curve for the air-filled guide) but then bend away from it towards the curve, $\frac{b}{a} = \frac{1}{2}$ (the guide completely filled with polythene). This is particularly evident on the curve $\frac{b}{a} = \frac{1}{8}$. The bend in the curves occurs where

$$\beta_2^2 = k_0^2 = k_1^2 - \left(\frac{\pi}{2b}\right)^2 \quad .. \quad .. \quad 2.5$$

which is the equation of a straight line through the origin in Fig. 3, and defines the frequency at which the wave in the air-filled part of the guide is plane and travels with the velocity of light in free space. It has been shown elsewhere[5] that the wave in the guide can be split up into two elementary plane waves travelling in directions making angles $\tan^{-1}\left\{\pm\frac{\sqrt{k^2-\beta^2}}{\beta}\right\}$ with the direction of the length of the tube. These angles are the critical or Brewster angles for the slabs of polythene when $\beta = k_0$, and above this frequency, which is defined in Equation 2.5, the elementary waves undergo complete internal reflection in the polythene. Below this frequency the transverse components of the electric and magnetic fields are at a maximum in the air-filled part of the guide, above it their maxima occurs in the dielectric.

## 3. Losses in the Dielectrics

In the preceding section, $j\beta$ was found as a first approximation to $P$ in Equation 2.3. In order to find the attenuation constant $\alpha$, let $k_1^2$ assume its correct value,

$$k_1^2 = \omega^2\mu_1\kappa_1 - j\omega\mu_1\sigma_1$$
$$= \omega^2\mu_1\left(\kappa_1 - j\frac{\sigma_1}{\omega_1}\right)$$

Then $P = \alpha + j\beta = j(\beta - j\alpha)$.

If $\sigma_1$ is sufficiently small, it may be assumed that $\beta$ has substantially the same value as for $\sigma_1 = 0$, and that $j\alpha$ is a small purely imaginary correction to it. If Equation 2.3 be written

$$f(\beta - j\alpha, \kappa_1 - j\frac{\sigma_1}{\omega}, \omega) = 0$$

then, since $j\alpha$ and $j\sigma_1/\omega$ are small, the function $f$ can be approximated to by the

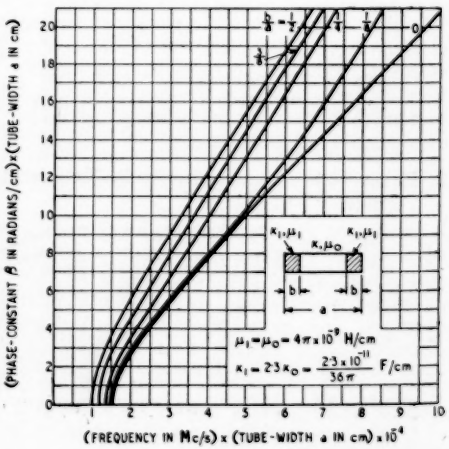


Fig. 3. *Phase-constant for $H_{01}$ wave with E parallel to interfaces between air and polythene.*

first two terms of its expansion in powers of $j\alpha$ and $j\sigma_1/\omega$. Thus

$$\left.\begin{aligned} f(\beta, \kappa_1, \omega) - j\alpha \frac{\partial}{\partial \beta} f(\beta, \kappa_1, \omega) \\ - j\frac{\sigma_1}{\omega}\frac{\partial}{\partial \kappa_1} f(\beta, \kappa_1, \omega) = 0 \end{aligned}\right\} \quad 3.1$$

and $f(\beta, \kappa_1, \omega) = 0$.

This reduces to

$$\frac{\alpha\beta}{\mu_1}\left[\frac{k_1^2 - k_0^2}{(k_0^2 - \beta^2)\sqrt{k_1^2 - \beta^2}} \operatorname{cotan}\{b\sqrt{k_1^2 - \beta^2}\} + \frac{\mu_1}{\mu_0}(\tfrac{1}{2}a - b) + b + \{(\tfrac{1}{2}a - b)\frac{\mu_0}{\mu_1}\frac{k_1^2 - \beta^2}{k_0^2 - \beta^2} + b\} \operatorname{cotan}^2\{b\sqrt{k_1^2 - \beta^2}\}\right]$$

$$= \tfrac{1}{2}\sigma\omega\left[b + b \operatorname{cotan}^2\{b\sqrt{k_1^2 - \beta^2}\} - \frac{\operatorname{cotan}\{b\sqrt{k_1^2 - \beta^2}\}}{\sqrt{k_1^2 - \beta^2}}\right] \quad .. \quad .. \quad .. \quad .. \quad 3.2$$

where $\beta$ is given by Equations 2.4.

Equation 3.2 gives the attenuation constant, $\alpha$, of an $H_{01}$-wave in the tube, the attenuation being due to the losses in the dielectric. Graphs of this attenuation, when the dielectric is polythene of conductivity $1.9 \times 10^{-6}$ mhos/cm, are given in Fig. 4 for several values of $\frac{b}{a}$. In every case, as the frequency increases indefinitely, the attenuation tends to the same value. The reason for this may be made clear from the discussion at the end of Section 2. As the frequency increases above the value, given by Equation 2.5, at which the wave travels down the guide with a phase-velocity equal to the velocity of light, more and more of the energy in the wave is confined to the solid dielectric in the guide. This means that a greater proportion of the energy in the wave is lost in heating up the dielectric. Thus for high frequencies the attenuation suffered by the wave in a partially-filled guide increases with increasing frequency, and in the limit has the same value as it would have if the solid dielectric completely filled the guide.

## 4. Losses in the Conducting Walls of the Tube

Up to this stage the walls of the tube have been assumed perfectly conducting. In fact, they have a finite, though very large, conductivity. The form of the wave is not appreciably distorted from its form in a perfectly conducting tube, but instead of perfect reflection occurring at the walls of the tube, a rapidly attenuated wave travels into them. The ratio of electric to magnetic field-intensity in this wave is given by[6]

$$\frac{E}{H} = \left(\frac{j\omega\mu_2}{\sigma_2}\right)^{1/2}$$

where $\mu_2$ is the specific inductance, and $\sigma_2$ the conductivity of the metal. The power going into this wave represents a loss of energy to the wave travelling down the tube.

The average power per cycle flowing through unit area in any direction is

$S$ = Real part of $\frac{1}{2}E \wedge H^*$,

where $E$ and $H$ are the electric and magnetic vectors perpendicular to the given direction,

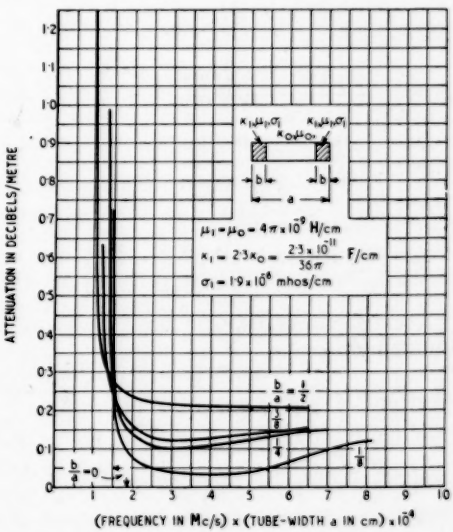
ATTENUATION IN DECIBELS/METRE
(FREQUENCY IN Mc/s) x (TUBE-WIDTH a IN cm) x 10⁴
μ₁ = μ₀ = 4π x 10⁻⁹ H/cm
κ₁ = 2·3κ₀ = 2·3 x 10⁻¹¹ / 36π F/cm
σ₁ = 1·9 x 10⁻⁶ mhos/cm
b/a = 0
b/a = 1/8
b/a = 1/4
b/a = 1/2

Fig. 4. *Dielectric losses for $H_{01}$ wave with E parallel to interfaces between air and polythene.*

$H^*$ denotes the complex conjugate of $H$ and $\wedge$ signifies vector product. Power flowing per unit area into the metal is, therefore,

$$S = \tfrac{1}{2}|H|^2\left(\frac{\omega\mu_2}{2\sigma_2}\right)^{1/2}$$

where $H$ is the tangential component of the magnetic-field intensity at the surface of the

tube and, provided the conductivity is large, is very nearly equal to $H$ of Equations 2.1 and 2.2 for walls of infinite conductivity. The small difference between them may be neglected.

Thus the average power absorbed per unit length of tube by the walls $z = \pm\frac{a}{2}$ is

$$S_1 = \frac{A^2 d}{\omega^2 \mu_1^2}\left(\frac{\omega\mu_2}{2\sigma_2}\right)^{1/2}(k_1^2 - \beta^2) \times \frac{\cos^2\{(\tfrac{1}{2}a - b)\sqrt{k_0^2 - \beta^2}\}}{\sin^2\{b\sqrt{k_1^2 - \beta^2}\}}$$

Average power absorbed per unit length by the walls $y = \pm\frac{d}{2}$ in Region I is

$$S_2 = A^2(\tfrac{1}{2}a - b)\left(\frac{\omega\mu_2}{2\sigma_2}\right)^{1/2}\frac{k_0^2}{\omega^2\mu_0^2}.$$

Average power absorbed per unit length by walls $y = \pm\frac{d}{2}$ in Region II is

$$S_3 = \frac{A^2 b}{\omega^2\mu_1^2}\sqrt{\frac{\omega\mu_2}{2\sigma_2}}\,k_1^2 \times \frac{\cos^2\{(\tfrac{1}{2}a - b)\sqrt{k_0^2 - \beta^2}\}}{\sin^2\{b\sqrt{k_1^2 - \beta^2}\}}$$

Average power flow in direction of tube through I is

$$W_1 = A^2 d\,\frac{\beta}{\omega\mu_0}\left[\tfrac{1}{4}a - \tfrac{1}{2}b + \frac{\sin\{(a - 2b)\sqrt{k_0^2 - \beta^2}\}}{4\sqrt{k_0^2 - \beta^2}}\right]$$

Average power flow in direction of tube through II is

$$W_2 = A^2 d\,\frac{\beta}{\omega\mu_1}\,\frac{\cos^2\{(\tfrac{1}{2}a - b)\sqrt{k_0^2 - \beta^2}\}}{\sin^2\{b\sqrt{k_1^2 - \beta^2}\}} \times \left[\tfrac{1}{2}b - \frac{\sin\{2b\sqrt{k_1^2 - \beta^2}\}}{4\sqrt{k_1^2 - \beta^2}}\right]$$

The attenuation constant due to losses in the conducting walls of the tube is[6]

$$\alpha' = \frac{1}{2}\cdot\frac{\text{Power loss per unit length of tube}}{\text{Power transmitted down the tube}} = \frac{S_1 + S_2 + S_3}{2(W_1 + W_2)}$$

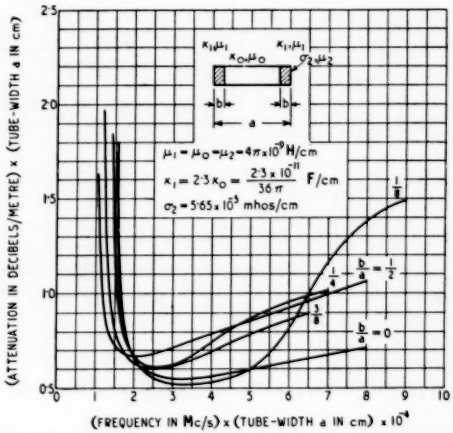
(ATTENUATION IN DECIBELS/METRE) x (TUBE-WIDTH a IN cm)
(FREQUENCY IN Mc/s) x (TUBE-WIDTH a IN cm) x 10⁻⁴
μ₁ = μ₀ = μ₂ = 4π x 10⁻⁹ H/cm
σ₂ = 5·65 x 10⁵ mhos/cm
b/a = 0
b/a = 1/8
b/a = 1/4
b/a = 1/2
b/a = 3/8

Fig. 5. *Copper losses for $H_{01}$ wave with E parallel to interfaces between air and polythene.*

Fig. 5 illustrates the attenuation for a copper pipe, of conductivity $5.65 \times 10^5$ mhos/cm. Again all the curves except that for $b = 0$ have the same asymptote as frequency increases indefinitely.

## REFERENCES


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[3] W. L. Barrow, "Transmission of Electromagnetic Waves in Hollow Tubes of Metal," *Proc. Inst. Radio Engrs.*, October 1936.

[4] S. A. Schelkunoff, "Transmission Theory of Plane Electromagnetic Waves," *Proc. Inst. Radio Engrs.*, November 1937.

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[6] Stratton, "Electromagnetic Theory," p. 543 *et seq.*

# COMPARISON OF AMPLITUDE AND FREQUENCY MODULATION

*By M. G. Nicholson*

*(Research Department, Colonial Radio Corporation, Buffalo, N.Y.)*

WHILE much has been published about the comparison of amplitude and frequency modulation, there is still much confusion on this controversial subject. Unfortunately, most of this confusion has been brought about by comparing the two under entirely different sets of conditions. The usual comparison has been between f.m. at 40–50 Mc/s with a channel width of about 200 kc/s, and a.m. restricted to a channel width of about 10 kc/s at a signal frequency of 1 Mc/s or less. Under such widely different conditions, static, natural and man-made, is radically different. Likewise, so called "high-fidelity" cannot be obtained on a restricted channel width of 10 kc/s, whether the type of modulation is f.m. or a.m. It is of interest to compare a.m. and f.m. under identical conditions of operation at the frequencies normally used for f.m. When the conditions of operation are substantially alike, and when good engineering is used in the design and construction of both systems, it has been found that there is surprisingly little difference in the final results.

This discussion is primarily centred around those conditions (frequency stability, channel-width, receiver bandwidth, etc.) normally realized in the v.h.f. band (30 to 300 Mc/s), as they affect interference and noise with f.m. and a.m. signals under similar conditions. Comparison of a.m. and f.m. will be made with respect to co-channel and adjacent channel interference, random noise, effects of frequency instability, and so-called "satellite" station performance.

## Interference

In the reception of two f.m. signals operating on substantially the same frequency, the weaker signal is suppressed to a great extent. In the case of two a.m. signals operating on substantially the same frequency, no such suppression exists. The resulting interference in a.m. is directly proportional to the ratio of the r.m.s. voltage of the weaker signal to the r.m.s. voltage of the stronger signal. Interference is defined here as the ratio of the r.m.s voltage in the a.f. output circuit which results from the presence of the weaker signal, to the r.m.s. voltage which results from full modulation of the stronger signal alone, [1,2]. Since the r.m.s. voltage of the weaker a.m. signal with 100 per cent modulation is only 3 db above that for no modulation, it follows that the interference is primarily that of the heterodyne beat note.

Theoretical and field test comparisons of a.m. and f.m. with the interfering signal on substantially the same frequency have been published,[3, 4, 5, 6] showing that under these restricted conditions f.m. will experience less interference than a.m. Unfortunately, virtually no information has been published concerning the comparison of a.m. and f.m. at frequency differences between the desired and undesired signals corresponding to adjacent-channel separation. Hans Roder points out that f.m. may be as much as 40 db inferior to a.m. under certain conditions of operation.[7] The degree of selectivity assumed by Roder in arriving at this value was definitely greater than is likely to be used. Calculations based upon Roder's method, but with a degree of selectivity more nearly that of an average f.m. receiver indicate marked superiority of a.m. over f.m. in the suppression of an undesired signal on an adjacent channel or on a second-channel separation.

In order to obtain more specific information on the relative abilities of f.m. and a.m. to combat interference, an f.m. receiver was used in making the comparison. For a.m. reception, the limiter grid circuit served as the a.m. detector, while the output of the limiter fed a balanced discriminator which served as the f.m. detector. The degree of selectivity in the receiver used in this test was quite similar to published data [8, 9] on receivers designed for f.m. using a 75-kc/s deviation. Curve A of Fig. 1 is the selectivity of this receiver. The bandwidth at an attenuation of 6 db is 175 kc/s. De-emphasis of the higher audio frequencies was applied to the outputs of the a.m. detector and of the f.m. discriminator, the

frequency response of the network being that of a 75-$\mu$sec time-constant circuit, which is the present recommendation for f.m. broadcast reception.

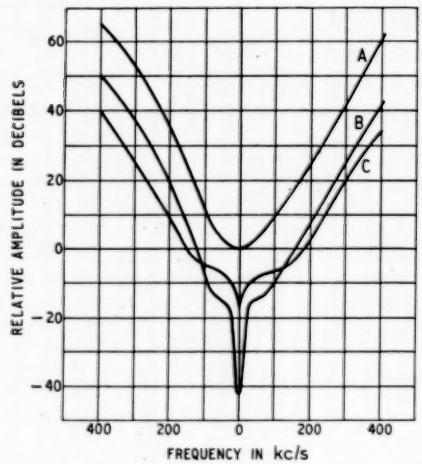



Fig. 1. *The selectivity curve of the receiver is shown at A, while B and C represent the signal levels causing 40 db interference for a.m. and f.m. respectively.*

The signal sources consisted of two standard-signal generators, one of which could be modulated with either f.m. or a.m. The a.m. was substantially void of f.m. since the modulation was accomplished in a stage well isolated from the oscillator. The f.m. was substantially void of a.m. as the result of careful design and adjustment of the modulator, as well as the action of the limiters following the modulator. This generator supplied the interfering signal for both f.m. and a.m. tests.

In this way, it was possible to use the same signal generators, and the same receiver for both f.m. and a.m., merely by selecting the type of modulation and the type of detection by simple audio-frequency switching. Besides being much more convenient, the accuracy of comparison is greater than if two different pairs of signal generators, and two different receivers had been used for completely separate f.m. and a.m. tests.

The degree of interference used in making these comparisons was such that the interference resulting in the a.f. output circuit was 40 db below the full modulation level of the desired signal. The modulation level of the undesired signal was adjusted to that level which produced maximum interference for each measured point, with the upper limit of modulation at 100 per cent for a.m. and 75 kc/s deviation for f.m. Full modulation produced maximum interference under all conditions of a.m., and also for all conditions of f.m. except where the desired signal and the undesired signal were sufficiently close together in frequency to cause an audible beat note. Under this latter condition, maximum interference occurred with little or no modulation on the undesired signal.

In the case of amplitude modulation the resulting interference is primarily that of the beat frequency between the carrier of the desired signal and the carrier of the undesired signal when the beat frequency is an audible frequency. If, however, the frequency of the beat is above audibility, the remaining interference is only that resulting from the modulation of the undesired carrier. F. E. Terman [10] points out that very great suppression of the weaker signal is brought about whenever the ratio of the signals at the detector is 2 : 1 or greater, provided a linear detector is used. This effect of a linear detector in suppressing the weaker signal has been discussed at some length in technical journals [7, 11, 12, 13, 14], but does not appear to be generally appreciated.

Curve B of Fig. 1 is the relative signal strength of an a.m. signal which interferes with a.m. reception to the extent of 40 db below the 100% modulation level on the desired a.m. signal. It is interesting to note that the vertical distance (db) between this curve and the selectivity curve (curve A) is substantially constant except in the small region where the frequency difference between the desired and undesired signals is within the audio-frequency range. In this region, the interference consists of an audible note in addition to the modulation of the undesired signal. Outside this region, the interference consists only of the modulation of the undesired signal. The degree of suppression of the modulation of the weaker signal by the stronger signal in this specific case amounted to about 23 db, since, in order to obtain a 40 db freedom from the modulation of the interfering signal, only about 17 db difference in signal voltages arriving at the linear detector was found necessary. This measured degree of suppression is in excellent agreement with that calculated for this case by the method used by Roder[7] in his comparison of a.m. and

f.m. for interference from a transmission on an adjacent channel.

Curve C of Fig. 1 is the relative signal strength of an f.m. signal which interferes with f.m. reception to the extent of 40 db below the full modulation level (75 kc/s deviation) on the desired f.m. signal. While the general shape of this curve resembles that of the selectivity curve (curve A), there is an increase in vertical distance (db) between these two curves as the frequency difference between the desired signal and the undesired signal is increased. This means that the degree of suppression of the undesired f.m. signal by the desired f.m. signal becomes less effective as the frequency difference is increased into the region where the selectivity of the tuned circuits is an appreciable factor in determining the ratio of the two signal voltages arriving at the input of the limiter.

By comparing Curves B and C of Fig. 1, it is seen that the degree of freedom from interference for f.m. is superior to a.m. only when the separation between the signals is substantially less than the 200 kc/s separation normally used in f.m. broadcast practice. Thus, interference from the two adjacent channels and the two alternate channels will be less with an a.m. system than with an f.m. system under the typical conditions used in these tests. This advantage of a.m. over f.m. is largely counter-balanced by the advantage of f.m. over a.m. where the frequency separation is less than 100 kc/s. However, if the separations between stations are made in multiples of 100 kc/s, instead of 200 kc/s, identical frequency assignments can be substantially eliminated without reducing the number of stations within an area. This results in less total interference with a.m. than with f.m.

### "Satellite" Station Operation

Experience in the v.h.f. band has shown that there are many localities which cannot be adequately covered by a single transmitter. The use of two or more a.m. stations on the same frequency carrying the same programme has been tried with rather poor results.

The use of two, or more, f.m. stations on the same frequency carrying the same programme has been somewhat more successful. This is primarily because of the so-called "capture" effect; that is, the stronger signal suppresses the weaker signal operating on the same frequency. Unfortunately, in those areas where the strengths of the two signals are almost equal, the limiter-discriminator combination is unable to receive either signal without considerable noise and distortion resulting from the mutual interference of one signal upon the other. This is often referred to as the "no-capture" area.

### "Diversity" Transmission

Recently there has been published[15] the results of a method of operating two or more transmitters on the same channel without the inherent defects of the f.m. or a.m. systems operating on a single frequency. This system employs two or three a.m. transmitters operating within a single v.h.f. channel but with the frequencies of the carriers separated by 20 or 30 kc/s. In this way there is no audible beat note. The phase of the audio modulation at each transmitter must be alike, in order that the audio components arising in the linear detector will likewise be in phase. Unlike f.m., there is a smooth, distortion-free, transition when going from the area where the signal from one transmitter predominates into the area where the signal from a second transmitter predominates. There is not even the slightest suggestion of a "no-capture" area. Of the more obvious applications for such a system is v.h.f. mobile communication, such as between Police Departments and moving cars, and for a public telephone service to privately owned automobiles where a greater service area than that obtainable from a single transmitter is required, for instance, over the complete length of a highway.

In addition to giving better general coverage, this system reduces materially the "flutter" to be found in v.h.f. systems when the receiver is being moved in a field consisting of reflections of the same magnitude as that of the direct wave. The reason why "flutter" is reduced so greatly is that the location of the points of low field intensity from one transmitter bear no relationship to the location of those from the other transmitter, thus the number of points where both field intensities are simultaneously at minima is drastically reduced. Thus, as long as one signal or the other is sufficiently strong, good reception will result, since there is no distortion brought about in the transition from one signal to the other. Where "flutter" is the

primary problem, the use of two a.m. transmitters with their aerials located as little as 100 feet apart will bring about a marked reduction in "flutter." This type of service has been called "diversity transmission" due to its similarity to diversity reception. There is one outstanding difference, which is that the additional equipment for diversity transmission is at the transmitter, while the receiver is unchanged. This is an important factor in mobile equipment where diversity reception is hardly practicable.

The question arises as to how practicable such a system would be for broadcasting. The most serious factor is the space-phase relationship; that is, the difference in arrival time at the receiver due to the difference in time paths from the common programme source. This would limit the spacing between stations to something of the order of ten miles. While such a system could be extended to cover any desired area with a single channel, by using many low-power transmitters linked together, it probably would be prohibitive in initial and maintenance expenses.

## Noise

Radio-frequency disturbances, generally referred to as "noise", consist of two distinctly different types. One type of noise produces "hiss" in the audio output, and is known variously as hiss, fluctuation, random, thermal agitation and "shot effect" noise. This fluctuation noise is characterized by a uniform distribution of energy with respect to frequency, with a ratio of peak voltage to r.m.s. voltage of about four for any given bandwidth of fluctuation noise.[16] The primary sources of fluctuation noise in a radio receiver are: (1) electron stream in the first stage of the r.f. amplifier, or in the frequency-changer if no r.f. amplifier is used; (2) electron thermal agitation in the aerial tuned circuits, and (3) fluctuation noise received by the aerial.

The second type of noise, impulse noise, consists of very short bursts of electrical energy, separated one from the other by periods of time substantially longer than the duration of an individual impulse. While impulse noise in the v.h.f. band may be caused by electrical storms (static), it is much more likely to be the result of man-made interference from electrical equipment such as electrical refrigerators, fans, elevators, switches, thermostatically controlled irons, oil-burning furnaces, automobile ignition systems, electric razors, or street cars.

## Fluctuation Noise

Frequency modulation has a distinct advantage over amplitude modulation in discriminating against fluctuation noise under most conditions of operation. The magnitude of this advantage varies with the conditions of operation. Also, care must be taken in interpreting the measured or calculated differences into their probable effect upon the listener, since the frequency distribution of the audible noise from an f.m. receiver is substantially different from that from an a.m. receiver.

It is quite generally agreed that the improvement against fluctuation noise by the use of f.m. as opposed to a.m. is an r.m.s. ratio of $\sqrt{3}$, or about 4.8 db, when the deviation is equal to the highest modulation frequency. Also, this improvement increases in direct proportion to the ratio of the deviation to the highest modulation frequency. For signal levels where the amplitude of the f.m. signal is less than twice the peak amplitude of the fluctuation noise, the above relationships are no longer true, but the r.m.s. difference between f.m. and a.m. is greatly reduced. This has been called the "sputter point." It is to be noted that this critical value of signal-to-fluctuation noise ratio, below which an f.m. receiver quickly deteriorates in performance, is dependent only upon the ratio of the signal-to-fluctuation noise, and not upon the ratio of the deviation-to-highest modulation frequency. This means that any increase in the bandwidth of the receiver, over and above that necessary just to accommodate the full deviation, will raise the critical value, thereby requiring a larger signal to produce the same performance as obtained with a minimum bandwidth receiver. This may be of significance when the bandwidth of the receiver is made considerably greater than the minimum in order to allow for frequency drift of the receiver oscillator as well as for inaccuracies in tuning.

For communication[17], where the transmission of intelligence is the primary objective, consideration of the low signal level condition is particularly important, since communication can be maintained at signal levels less than twice that of the peaks of the fluctuation noise. If it is possible to maintain proper tuning when using a band-

width only sufficiently wide to accommodate a deviation equal to the highest audio modulation frequency, f.m. should have a greater range than a.m. If, however, the bandwidth must be substantially wider, it becomes doubtful if f.m. will have any advantage over a.m. in discriminating against fluctuation noise in a communication system.

There now is a definite trend in the other direction; that is, towards the design of f.m. communication receivers with the bandwidth substantially less than that of the maximum deviation of the transmitter. In order to receive full deviation, sufficient extra amplification and limiting is included so that the attenuation of the carrier at maximum excursion due to the selectivity is compensated for in the limiter. This will work satisfactorily at any level of signal substantially above the fluctuation noise level. However, at low signal levels, a parodoxical situation exists; that is, the signal-to-noise ratio, as conventionally measured may be quite good, yet the intelligibility of the received signal may be very poor. Under these conditions of test, f.m. would appear decidedly superior to a.m., yet field tests show little difference in performance where intelligibility is the criterion.

This apparent discrepancy is readily understood if the action of such an f.m. receiver at signal levels comparable with that of the fluctuation noise is carefully analysed. As an example, let the signal be twice as strong as that signal which is just at the sputter point. Now, by slowly detuning the receiver from the carrier, the fluctuation noise will rise slowly until the sputter point is reached, beyond which, the noise quickly rises to the level which the receiver has in the absence of a carrier. Since the level of the carrier is only double that for the sputter point at exact tuning, the degree of detuning to reach the sputter point corresponds to two-to-one down on the selectivity curve of the receiver. Instead of detuning the receiver, frequency modulating the transmitter will, of course, do the same thing, except that it will be done at the modulating frequency. Thus, when the deviation extends beyond the sputter points, on each side of the centre, the modulation will be greatly distorted, and during those parts of the modulation cycle beyond the sputter point, noise will be, in effect, substituted for the modulation. From this it is seen that the use of high selectivity can result in an excessive rise in noise with modulation and high distortion at low signal levels, while at medium signal levels the performance is entirely normal.

Probably the reason for designing f.m. receivers with excessive selectivity lies in the method used for measuring the signal-to-noise ratio of a receiver. This method yields better apparent signal-to-noise ratios as the selectivity is increased. The method consists in comparing the noise output of a receiver, when there is no modulation, with the total output of the receiver at full modulation. If this is carried to the extreme, the output during modulation consists primarily of noise, modulated at the modulation frequency; thus, the so-called signal-to-noise ratio can be the noise-to-noise ratio, where the noise during modulation is compared to the noise when there is no modulation.

## Pre-Emphasis and De-Emphasis

A closer examination of the nature of the audible noise arising from fluctuation noise in f.m. and a.m. receivers reveals marked differences[5, 18]. The audible noise from an a.m. receiver has the same general characteristic as fluctuation noise; that is, equal noise energy per unit of frequency increment. On the other hand, the audible noise in an f.m. receiver (without special provisions) does not have this flat frequency characteristic, but the noise energy per unit of frequency increment increases as the square of the audio frequency corresponding to each increment. Since the higher audio frequencies have the ability of arresting one's attention so much more than the lower audio frequencies, it becomes almost imperative to do something to reduce them. This is normally done by " tone controlling " the receiver, now known as de-emphasis, while at the transmitter an inverse network (so-called pre-emphasis) is used to accentuate the higher audio frequencies. In this way, the overall response of the system is made uniform. This addition to the f.m. system improves its signal-to-noise ratio. The addition of pre-emphasis and de-emphasis to an a.m. system will likewise improve its signal-to-noise ratio. However, the r.m.s. value of improvement is greater for f.m. than for a.m., since most of the noise in the f.m. receiver is in the high end of the audio spectrum, as compared to a uniform distribution in the a.m. receiver. The use of de-emphasis in an f.m. receiver will alter the noise energy distribution so that it is flat above that frequency at which the

de-emphasis network becomes effective. While for an a.m. receiver, the noise energy distribution remains flat up to the point where the de-emphasis network becomes effective, above which, the noise energy per unit of frequency increment decreases at the rate corresponding to the inverse of the square of the frequency.

From this, one must conclude that while f.m. is, in general, superior to a.m. in discriminating against fluctuation noise, it is necessary to take into account the effect of different noise distribution spectra upon the listener as a correction factor upon any calculated or measured r.m.s. value of comparative merit. This means that the effectiveness of f.m. over a.m. against fluctuation noise is less than that indicated by the r.m.s. values. However, in addition to the noise distribution correction, there is another factor which is brought about by a necessary reduction of programme level entering the pre-emphasis network at the transmitter. This must be done in order to avoid over-modulation, since there is an appreciable amount of energy in the higher frequency part of the audio spectrum. Published calculations on the improvement brought about by the use of pre-emphasis and de-emphasis have usually ignored this correction, by assuming that the pre-emphasis can be applied without a reduction in the level of the lower audio frequencies.

## Standard Pre-Emphasis

The degree of pre-emphasis for f.m. broadcasting has now been standardized as the voltage—frequency characteristic of an inductor and resistor in series, with a time constant of 75 $\mu$sec, and being fed from a constant-current source. Published data on the previous standard of 100 $\mu$sec time constant of pre-emphasis gave a theoretical improvement of 7.4 db to f.m. over a.m., when the programme modulation level correction was not taken into account.[5] The necessary reduction in programme level entering the 100-$\mu$sec pre-emphasis network has been reported as being from 2.5 db to 4.5 db for various types of programme material.[18]

The British Broadcasting Corporation reports[19] that a reduction of 6.5 db in programme level is generally necessary for 100-$\mu$sec pre-emphasis. While for a 50-$\mu$sec pre-emphasis the necessary level reduction is reported as 3 db. Whatever may be the magnitude of level reduction necessary for 75-$\mu$sec pre-emphasis, it should be taken into account when calculating the advantage of pre-emphasis.

## Impulse Noise

Impulse noise seldom causes perceptible interference, unless the peak amplitudes of the impulses are equal or greater (usually considerably greater) than those of the signal being received, whether it is an f.m. or an a.m. signal. During these very brief periods the noise obscures the signal, while in the relatively long period between the impulses, the signal is free from noise. While the individual impulses differ greatly in duration, depending upon the electrical characteristics of the source, they are almost always substantially less than one microsecond in length, prior to passing through tuned circuits.

Practically all of the published articles dealing with the theoretical considerations of the effect of impulse noise upon an f.m. receiver in comparison with that upon an a.m. receiver have assumed the peak voltage of the noise impulses to be, at a maximum, equal to that of the signal voltage. Under these restricted conditions, f.m. is theoretically superior to a.m. in rejecting noise impulses, but in practice, such low-level noise impulses are insignificant. Probably the reason for the confinement of the theoretical considerations to this region is that the mathematical processes used have been applicable only to this restricted region of operation, and become discontinuous functions as soon as the peak voltage of the noise exceeds that of the signal. This is most unfortunate for it is the effect of high level impulse noise that is the primary factor in comparing a.m. with f.m. against impulse noise as normally encountered in practice.

The duration of noise impulses is much too short to pass through the tuned circuits of the receiver without being lengthened considerably. The final lengths of such short impulses after passing through the tuned circuits are almost entirely independent of their original lengths, and are dependent rather upon the bandwidth and phase characteristics of the receiver.[16, 20]

The final length of an impulse is inversely proportional to the bandwidth of the receiver, while the peak amplitude is directly proportional to bandwidth. Thus, the greater the bandwidth of the receiver, f.m. or a.m., the shorter will be the time during which a noise impulse will obliterate the signal. Also, the

greater will be the time during which the signal can be received in those periods of time between noise impulses. Since the amplitude of the impulses increases with an increase in bandwidth, it becomes necessary to use some sort of amplitude limiter, in both f.m. and a.m. receivers, in order to realize the advantage brought about by the use of increased bandwidth.

The action of a strong impulse upon such a receiver is to put a "hole" in the signal, lasting perhaps ten microseconds for a receiver with bandwidth characteristics such as the one used in the example given in the interference measurements, and shown in Fig. 1, Curve A. If such impulses occur at a rate of 1,000 per second, they should interfere with the signal only 1 per cent of the total time, while the signal should be received 99 per cent of the total time with no interference.

**Radar Conditions**

A very good example of a highly sensitive receiver operating in the presence of strong impulse interference is a radar receiver. The interfering impulses are the transmitting impulses received either directly from the common feed line to the common aerial for transmitting and receiving or from a receiving aerial located only a few feet from the transmitting aerial, while the desired signals are the reflected waves arising from the transmitted waves being reflected back to the receiver. Here the interfering signal is likely to be several hundred volts at the input terminals of the receiver, while desired signals (echoes) of only a few microvolts may be received with substantially no interference during the time between transmitting impulses (interference).

It is interesting to note that the interference voltage is often more than a million times greater than the desired signal voltage, yet substantially no interference exists except during the short periods of time when the transmission of the impulses is taking place.

Three important factors affecting the performance of a receiver (f.m., a.m. or radar) in the presence of impulse interference are the following:

(1) The time necessary for the receiver to return completely to its condition prior to the reception of a noise impulse.

(2) The increase in length of an impulse in passing through the receiver.

(3) The amplitude of the impulse after passing through the receiver in comparison with that of the desired signal.

While the recovery time (1) is dependent upon the effective bandwidth, it may be adversely affected by the use of a circuit whose operating conditions are changed by the reception of an impulse and do not quickly return to their original state. An example of this is a grid circuit consisting of a coupling capacitor and a grid resistor, whose time constant is appreciably greater than the duration of the impulse. An impulse with a positive amplitude greater than the negative grid bias voltage will cause grid current to flow, thus changing the charge in the capacitor, resulting in an increased bias by the end of the impulse. This increase in negative bias may be sufficient to render the tube inoperative for an appreciable length of time after the impulse. Under extreme conditions the impulses may occur sufficiently often to keep the tube beyond cut-off between impulses, thus eliminating the desired signal entirely. One remedy for this could be the decrease of the time constant of the circuit to the point where the circuit could follow the wave shape of the impulse. Another remedy could be the increase of the value of capacitance to a sufficiently high value such that a single impulse would not cause an appreciable change in the voltage across the capacitor, and the reduction of the value of resistance in the grid circuit to the point where the summation of impulses would not raise the average level of grid bias appreciably.

The increase in impulse length (2) brought about by the receiver is primarily dependent upon the effective bandwidth of the receiver. In order to obtain minimum impulse noise interference for either f.m. or a.m., the bandwidth prior to the limiter should be made as wide as possible, consistent with other receiver requirements. Also, the effective bandwidth after the limiter should be as narrow as possible, consistent with other receiver requirements [21, 22].

The amplitude of the impulse (3) after passing through the receiver is dependent upon the effectiveness of the limiter and the effect of the limited amplitude upon the circuits following the limiter.

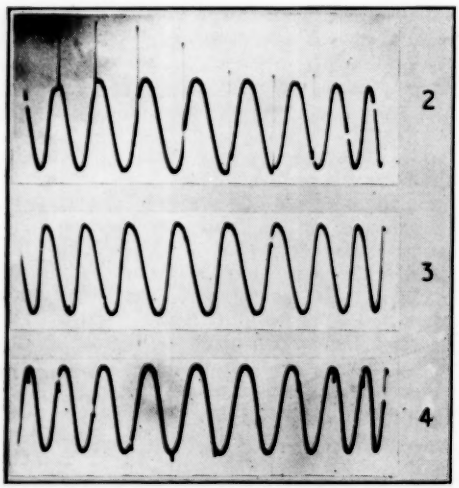
Up to the point of limiting, there is little difference in the basic design of a receiver for minimum impulse interference suscep-

tibility, whether it is for f.m. or a.m. The limiter and detector for f.m. are quite different in design and action from those for a.m. Due to these large differences, each system will be described separately, with the comparisons confined to the final results of each system.

**Impulse Noise in F.M. Reception**

The combination of an amplitude limiter and a balanced discriminator, operating in this order, is used for f.m. throughout these comparisons[23]. The action of the limiter is such that the equivalent frequency of its output is substantially that of the stronger signal, whether it is the desired signal or a noise impulse reaching the input of the limiter; the amplitude of the output is a constant.

First, let us consider the reception of an unmodulated carrier. The steady output voltage of a balanced discriminator will be dependent upon the relationship between the carrier frequency and the discriminator balance frequency. The application of a carrier with a frequency slightly higher than the balance frequency will cause the discriminator to have a finite output voltage, but, if the applied carrier has a frequency slightly lower than the balance frequency, the output voltage will be of opposite polarity. Over normal operating range of modulation (up to maximum frequency deviation) this relationship should be linear in order to be able to convert the frequency modulated wave into the corresponding audio-frequency wave without introducing distortion.



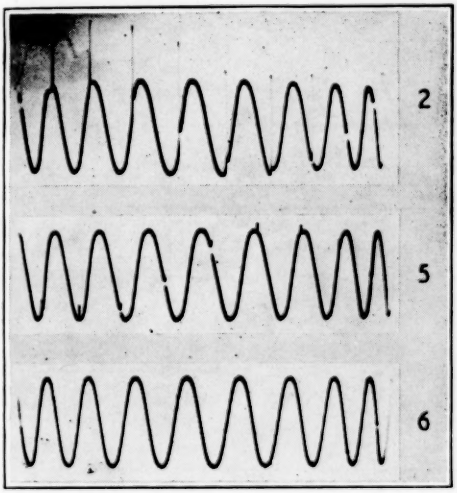

Figs. 2, 3 and 4. *Impulse noise in a.m. reception without a limiter; with an accurately tuned f.m. receiver; and with an f.m. receiver detuned by 50 kc/s.*

Next, let us consider the reception of strong noise impulses having the usual short duration, but unaccompanied by a signal carrier. Since the duration of the impulses is too short to be passed without being considerably lengthened by the effect of the tuned circuits, it follows that the average or equivalent frequency of the impulse arriving at the limiter is the centre frequency of the selectivity response curve, provided it is symmetrical. Likewise, the equivalent frequency of the output of the limiter during the impulse is this centre frequency, but with a limited amplitude. For the noise impulse to produce minimum disturbance, the balanced discriminator must be aligned to balance at this frequency. If, however, the centre frequency of the selectivity curve does not coincide with the frequency at which the discriminator balances, noise impulses will reach the audio amplifier. The amplitude of the impulse from the discriminator for misalignment corresponds to the peak output voltage at this same point as produced by an f.m. signal having a deviation equal to the magnitude of this misalignment.

Now, let us consider the reception of an unmodulated carrier accompanied by noise impulses. Since the equivalent output frequency of the limiter is determined by whichever is the stronger at any instant, it follows that output of the discriminator alternates between the values determined by the carrier alone and by the noise impulse when it is in control of the limiter. If the output voltage and polarity are the same for the two conditions, there will be practically no noise. If, however, the output voltages are not the same, noise impulses will be reproduced with an amplitude corresponding to this difference in voltage. Thus, if the receiver is detuned 25 kc/s from the point of minimum impulse interference, the amplitude of the impulses at the output of the discriminator will be the same as the peak value obtained from an f.m. signal having a deviation of 25 kc/s. Also, if the carrier is frequency modulated, the peak value of a noise impulse will be dependent upon the instantaneous frequency of the signal at the

time of arrival of the noise impulse. It should also be noted that the tuning point for minimum impulse interference is not the discriminator balance point unless the selectivity ahead of the limiter is symmetrical.

The peak amplitude of the noise impulse can be reduced by passing the output of the discriminator through a low-pass filter designed to eliminate the super-audible frequencies, yet still pass the audible frequencies. The degree of reduction in peak amplitude of such impulses will be the ratio of the effective bandwidth of the receiver to the limiter input divided by the bandwidth of the low-pass filter. The use of 75-$\mu$sec de-emphasis will produce this effect to the extent of about eight-to-one for a typical f.m. receiver. This same ratio of eight-to-one brought about by de-emphasis will also be true for an a.m. receiver having similar wide-band responses through the detector but with the restricted frequency response in the a.f. amplifier.




Figs. 2, 5 and 6 (above). *Impulse noise in a.m. reception without a limiter; with a Kaar-Fyler limiter, and with the limiter of Fig. 8.*

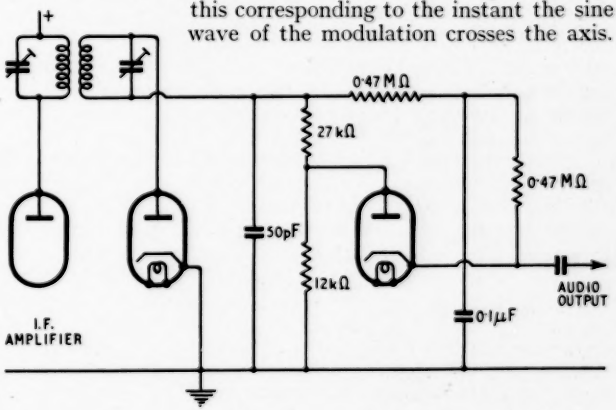



Fig. 7 (right). *The Kaar-Fyler noise limiter.*

Oscillograms of noise impulses mixed with the signal under various conditions of operation show the action of noise limiting. Fig. 2 is an oscillogram showing the relationship of the noise impulses to the modulation used in the oscillograms to follow. The modulation is a sine wave of one frequency, while the repetition rate of the noise impulses corresponds to a pulse per cycle of a frequency differing slightly from the modulation frequency. This enables one to obtain a single oscillogram of several cycles showing the results of noise impulses occurring at various parts of the audio cycle. This oscillogram was taken with a.m. instead of f.m., but at an identical signal level with the f.m. oscillograms to be described. The same receiver was used for both f.m. and a.m. The limiter input grid was used as the diode for a.m. reception, thus making it possible to have identical circuit conditions for both f.m. and a.m. ahead of the f.m. limiter and the a.m. detector. The same degree of de-emphasis was used for both. The levels of modulation were 30% for a.m. and 30% of the maximum allowable 75-kc/s deviation for f.m.

Fig. 3 is the oscillogram for f.m. accurately tuned to give minimum noise at zero deviation. The magnitude of the resulting impulse is practically zero if the impulse occurs at zero deviation of the modulating frequency, this corresponding to the instant the sine wave of the modulation crosses the axis.

The result of detuning the receiver by 50 kc/s is shown in Fig. 4. The polarity of the resulting impulses is dependent on which way the receiver is detuned from the discriminator balance frequency. A comparison of Figs. 3 and 4

indicates the importance of exact tuning for maximum rejection of impulse noise.

### Impulse Noise in A.M. Reception

Fig. 2 is the oscillogram of the a.m. detector output after passing through a de-emphasis network equivalent to the one used between the f.m. discriminator and the audio output. If no de-emphasis were used, the amplitude of the impulses would be several times larger, being limited to the point of overload of the i.f. amplifier of the receiver. Thus, the peaks of all the impulses would have the same overload level. However, the de-emphasis network reduces the inadvertently limited impulses by a relatively constant ratio, resulting in the impulses appearing to vary in amplitude with the instantaneous value of the modulation at the time of occurrence of each of the impulses. Attempts to photograph the trace obtained with no de-emphasis were unsuccessful, due to the low intensity of the trace resulting from the high velocity of the spot on the oscilloscope screen during the impulse.

Fig. 5 is the oscillogram of the output of an a.m. detector after passing first through an amplitude limiter and then through the de-emphasis network. Any one of several types of limiters could have been used to give comparable results [21, 24, 25]. The particular limiter used in this receiver consisted of a diode in series with the output of the detector, so connected and biased as to operate as a closed circuit for any amplitude of signal up to that equivalent to 100 % modulation, but to become an open circuit to greater amplitudes. This particular arrangement is often referred to as "series diode" or Kaar-Fyler[26] noise limiter. The circuit and values used in these tests are shown in Fig. 7.

Fig. 6 shows the results obtained with a.m. in conjunction with a noise limiter developed by the writer. The operation of this system of noise limiting is to leave the instantaneous audio voltage at whatever voltage it was at the beginning of a noise impulse until the noise impulse is over. If the duration of the noise impulses is short compared to the length of a cycle of modulation, the impulses will cause only very slight disturbances on the modulation, since at the end of the noise impulse the instantaneous audio voltage needs to shift only a very small amount to be back to the audio output voltage of the detector.

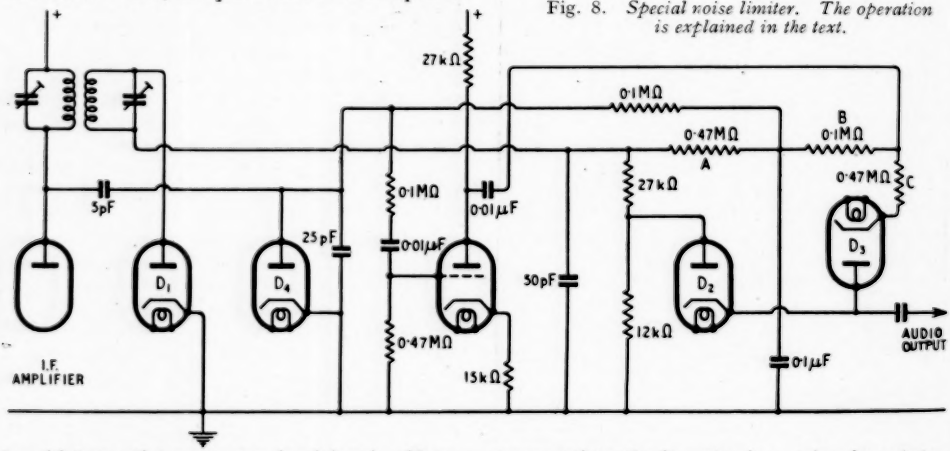



Fig. 8. *Special noise limiter. The operation is explained in the text.*

The action of these a.m. receiver noise limiters theoretically is not adversely affected by detuning so long as the signal is kept within the pass-band of the receiver. In the receiver used detuning the signal by 50 kc/s had no discernible effect upon the resulting signal-to-noise ratio or upon the oscillograms of the signal and noise impulses.

Fig. 8 is a circuit diagram of this special noise limiter used in obtaining Fig. 6. Diode 1 serves as the detector. The audio output is taken from a tap on the diode load through Diode 2. Diode 2 is made conductive by passing through it a polarizing current derived from Diode 1. The path of this current is through resistors A, B and C and Diode 3. The 0.1-$\mu$F capacitor at the junction of A and B serves to attenuate the

audio-frequency component to a negligible value. Diode 4 is biased by the voltage at this junction so as to make it inoperative on signals with peaks equal to or less than twice the value of the carrier. For high amplitude noise impulses it becomes operative, with the detected impulses being amplified by the triode and impressed upon Diode 3 with such a polarity as completely to counteract the normal polarizing current. Thus, Diode 2 and Diode 3 are effectively open circuits for the duration of an impulse. At the completion of an impulse, normal operation is quickly resumed with the instantaneous voltage of the audio output having been only slightly displaced by the action of the impulse. Such a system as this should give better rejection of impulse noise in a.m. than that to be expected in f.m. even with most favourable adjustments for f.m.; that is: (1) symmetrical noise impulses over the bandwidth of the receiver, (2) symmetrical band-pass characteristics of the receiver, (3) perfect limiting, and (4) perfect tuning. Practical results, when extreme care is taken to fulfil these exacting requirements for f.m., substantiate the conclusion that a.m. is superior to f.m. in rejecting impulse noise, as can be noted by comparing Figs. 6 and 3.

## Impulse Noise Generation

The noise impulses used in these comparisons were generated by spark discharges. The discharge was accomplished by charging a capacitor to a sufficiently high voltage to cause a spark gap connected in the circuit to break down. A 10-ohm resistor was connected in series with the spark gap. The impulse noise voltage was taken from across the resistor, passed through an attenuator and then through a shielded cable to the receiver under test. The complete generator was sufficiently well shielded to reduce any stray radiation to a negligible value. The duration of the impulse was substantially less than one microsecond.

Noise impulses from two other types of generators were compared with those from the spark gap generator, with varying results, but in general even less favourable results were obtained for f.m. in comparison with a.m. than with the spark gap impulse generator.

One of the other types of noise generator consisted of a pulse-modulated radio-frequency amplifier and a standard-signal generator. The signal leakage through the modulated amplifier during the time between impulses actually caused more interference than the impulses themselves, even though this leakage was only about one per cent. This equipment did not give any true indication as to the impulse noise susceptibility of either f.m. receivers or a.m. receivers.

The remaining type of noise generator tested in these comparisons consisted of an oscillator and an impulse generator so arranged that the oscillator operated only during an impulse; thus, the signal could be received with no interference during the time between noise impulses. Oscillograms of f.m. and a.m. reception through the impulse noise interference generated in this manner were substantially identical with those obtained with the spark-gap generator, provided the duration of the impulse was not over 5 μsec. For longer impulses, extreme care had to be taken in aligning the interfering signal centre frequency to that of the signal carrier, otherwise the f.m. performance was poorer than with the spark-gap interference. This is to be expected when one considers the characteristic energy distribution of an impulse in respect to its duration[27]. As an example, for a duration of a radio-frequency impulse of 2 μsec, with constant amplitude during the impulse, the relative amplitudes of the various components within 200 kc/s of the central frequency are shown by Curve A of Fig. 9. This curve represents the locus of the amplitudes of these various components, while the spacing between the

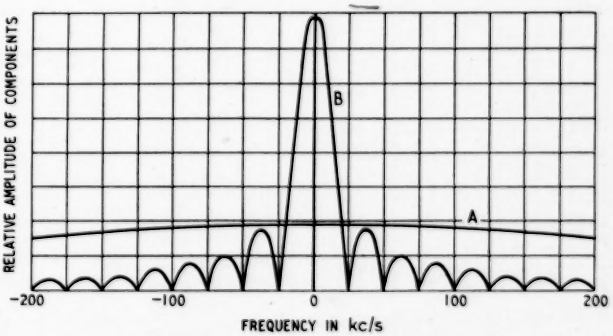



Fig. 9. *Curves A and B show respectively the amplitudes of components of 2-μsec and 40-μsec impulses.*

individual components is dependent only upon the repetition rate of the impulses. If, for example, the repetition rate were 1,000 per second, there would be a spacing of only 1 kc/s between the adjacent components.

Since there is so little difference in the amplitudes of the various components of a 2-$\mu$sec impulse around the centre frequency, it follows that small discrepancies in the alignment of this centre frequency to that of the desired signal will have practically no effect upon measurements of signal-to-noise ratios.

If, on the other hand, one attempts to measure impulse susceptibility of an f.m. receiver by applying relatively long duration impulses, such as 40$\mu$sec, very accurate alignment is required, otherwise erratic readings will be obtained. It becomes apparent why this is the case when one analyses the spectrum resulting from 40-$\mu$sec impulses. Curve B of Fig. 9 represents the locus of the amplitudes of the various components of 40-$\mu$sec impulses. Since this curve is much " sharper " than the selectivity of the receiver being tested, the equivalent frequency of the output of the limiter will be substantially equal to the centre frequency of the impulses. If this frequency differs from the frequency of the f.m. signal, the output of the discriminator will contain impulses corresponding in peak amplitude to that resulting from an f.m. signal with a deviation equal to this difference in frequency. Perfect limiting and re-tuning of the receiver will not improve the results, since the equivalent output frequency of the limiter is shifting to and fro between that of the carrier and that of the noise impulses. However, if 2-$\mu$sec impulses were used instead of the 40-$\mu$sec impulses, the equivalent frequency of the impulse output of the limiter would be essentially that of the centre of the selectivity curve since those components passing through the receiver have substantially equal amplitudes upon entering the receiver; thus, proper tuning of the receiver will give minimum impulse noise response, as in actual practice.

## Conclusions

F.M. is superior to a.m. only where fluctuation noise is the limiting factor.

A.M. is superior to f.m. in the matter of interference, even though the selectivity of the a.m. receiver is identical with that of the f.m. receiver. The optimum station spacing for a.m. is somewhat different from the 200-kc/s spacing used in f.m.

A.M. is superior to f.m. in discriminating against impulse noise.

A.M. is much less adversely affected by imperfect tuning than is f.m.

A.M. is superior to f.m. in " satellite " station operation.

In a given area (city, state, or country) it would appear that for a given band in the v.h.f. spectrum, more signals would be available to any individual receiver (with the same bandwidth for f.m. and a.m.) by the use of a.m. provided sufficient additional transmitting power were available to overcome the fluctuation noise problem in fringe areas, or sufficient additional stations were used to reduce substantially the size of the fringe area.

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# VERY-WIDE BAND RADIO-FREQUENCY TRANSFORMERS

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*(Concluded from page 177 June issue)*

## 5. Low-pass Filter Technique

WE have shown methods for the determination of $L$ and $C$. They are both of them a nuisance as they restrict the value of $f_2$. We can, however, make the best of a bad job by arranging that together they form a single $\pi$-section low-pass filter. This need only be done when the frequency range required is rather great for the desired impedance level. In most cases it is only necessary to make $L$ as small as possible and leave $C$ to look after itself. The transformer may then be looked upon at the high-frequency end of the pass range as an ideal transformer.

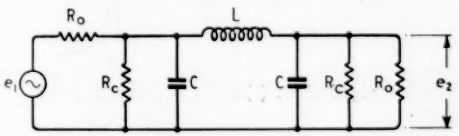



Fig. 17. *Low-pass filter equivalent of a transformer.*

Consider now a low-pass filter formed by $L$ and $C$, Fig. 17. $R_c$ represents the shunt losses at $f_2$, usually due to dielectric losses in the interwinding insulation. $R_c$ can usually be left out of account if good insulation is used, but becomes of importance at impedance levels higher than about 500 ohms at frequencies greater than about 25 Mc/s. If we let $R_c \to \infty$ and

$$(2\pi f_2)^2 LC = 1 \qquad .. \quad .. \quad .. \quad (32)$$

and $$2\pi f_2 L = R_0 \qquad .. \quad .. \quad .. \quad (33)$$

then the amplitude and phase responses of $\frac{e_2}{e_1}$ are shown in Fig. 18 (*a*). If we let $R_c$ have such a value that the insertion loss is 2 db at $f_2$ we obtain Fig. 18 (*b*). The value of $R_c$ which brings this about is

$$R_c = 6.33 R_0 \qquad .. \quad .. \quad .. \quad (34)$$

A technique much employed is to make

$$L = \frac{R_0}{2\pi f_2} \qquad .. \quad .. \quad .. \quad .. \quad (33)$$

and arrange that

$$C < \frac{1}{2\pi f_2 R_0} \qquad .. \quad .. \quad .. \quad (35)$$

and then add the necessary fixed capacitances on to the $C$'s in order to raise the total terminating capacitance to the correct value. Quite often it is necessary only to ensure that both $C$ and $L$ obey the inequalities (35) and (36).

$$L < \frac{R_0}{2\pi f_2} \qquad .. \quad .. \quad .. \quad .. \quad (36)$$

This means that the actual cut-off frequency will be greater than $f_2$. This cannot be allowed, however, if the transformer is required to discriminate against frequencies outside the stated range.

The information so far given is sufficient to design a normal two-winding transformer; but before giving any example of actual design we shall discuss some further ques-

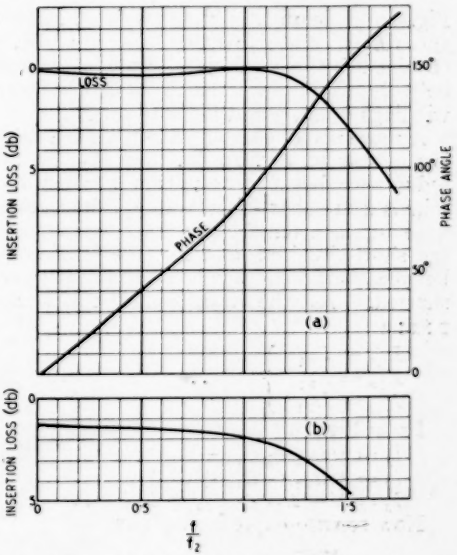



Fig. 18. *Amplitude and phase responses of low-pass filter (a) when $R_c \to \infty$ and (b) when $R_c$ is of such value that the insertion loss is 2 db at $f_2$; i.e., $R_c = 6.33\ R_0$.*

tions, some of which are directly applicable only to multi-winding transformers. The insight gained by the following discussion will be of value, however, in the design of any transformer.

## 6. Partial Leakage Inductance

Consider a three-winding transformer with one end of all windings at a common potential and of impedance ratios one to one to one, Fig. 19 (*a*). Neglect for simplicity all shunt losses, including the shunt capacitances. Neglecting the shunt capacitances will in no way affect the following discussion because they can always be put across the transformer windings afterwards. If we re-draw

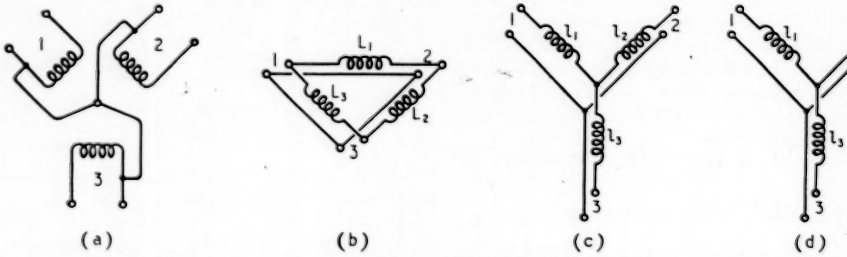



Fig. 19. *A three-winding transformer* (*a*) *has the delta and star equivalents* (*b*) *and* (*c*). *The last can be reduced to* (*d*).

Fig. 19 (*a*) leaving only the leakage inductances $L_1$, $L_2$ and $L_3$ we arrive at (*b*) which is obtained by regarding $L_1$ as the leakage between windings 1 and 2, $L_2$ as that between windings 2 and 3 and likewise for $L_3$.

A more useful combination than the delta is the star of Fig. 19 (*c*). Let $L_{12}$ be the inductance between terminals 1 when terminals 2 are short-circuited, $L_{13}$ the inductance between terminals 1 when terminals 3 are short-circuited, and $L_{23}$ the inductance between terminals 2 when terminals 3 are shorted. As the transformer ratios are 1 : 1 : 1

$$\left.\begin{aligned} L_{12} &= L_{21} \\ L_{13} &= L_{31} \\ L_{23} &= L_{32} \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (37)$$

From Fig. 19 (*c*)

$$l_1 = \tfrac{1}{2}(L_{12} + L_{13} - L_{23}) \quad .. \quad (38.1)$$
$$l_2 = \tfrac{1}{2}(L_{12} - L_{13} + L_{23}) \quad .. \quad (38.2)$$
$$l_3 = \tfrac{1}{2}(-L_{12} + L_{13} + L_{23}) \quad .. \quad (38.3)$$

Now rewrite expression (8) as

$$L = \frac{4\pi n^2}{1{,}000} \cdot \frac{A}{h} \quad .. \quad .. \quad .. \quad (39)$$

where $A$ is the cross-sectional area of the annular space between any two spiral windings of wide, infinitely thin copper strip. Putting (39) into (38)

$$l_1 = \frac{2\pi n^2}{1000h}\left(A_{12} + A_{13} - A_{23}\right) = \frac{4\pi n^2}{1000h} A_{12} > 0 \quad .. \quad (40.1)$$

$$l_2 = \frac{2\pi n^2}{1000h}\left(A_{12} - A_{13} + A_{23}\right) = 0 \quad .. \quad (40.2)$$

$$l_3 = \frac{2\pi n^2}{1000h} \cdot \left(-A_{12} + A_{13} + A_{23}\right) = \frac{4\pi n^2}{1000h} A_{23} > 0 \quad .. \quad (40.3)$$

Thus Fig. 19 (*c*) becomes (*d*).

Suppose that the middle winding is spaced equally from the inner and outer windings, $A_{12} = A_{23} = A$. Now parallel terminals 1 and 3. The result is a two-winding transformer of leakage inductance only half what it would have been had we wound a simple two-winding transformer with the same spacing between windings as was used with the three-winding transformer. We would gain no such advantage by paralleling terminals 2 and 3. If we now assume that our transformer consists of three *interwound* spiral windings of infinitely thin wide copper strip, we get from equations (9) and (38)

$$l_1 = \frac{4\pi}{1000h}\left[nA + 4\pi(n+2)(n-1)\epsilon^2\right] > 0 \quad .. \quad .. \quad (41.1)$$

$$l_2 = -\frac{8\pi^2}{1000h}(n+2)(n-1)\epsilon^2 < 0 \quad (41.2)$$

$$l_3 = \frac{4\pi}{1000h}\left[nA + 4\pi(n+2)(n-1)\epsilon^2\right] = l_1 > 0 \quad .. \quad (41.3)$$

where $\epsilon$ is the distance between turns in centimetres and $A$ is as before in square centimetres. If, as before, we connect

terminals 1 in parallel with terminals 3 and use the device as a simple two-winding transformer, the leakage inductance becomes $\frac{1}{2n}$ of the value it would have had in an ordinary two-winding transformer with the same physical constants but wherein the windings are not interwound. An interwound two-winding transformer would have $2\left[1+\frac{2\pi\epsilon^2}{nA}(n+2)(n-1)\right]$ times the leakage inductance of the interwound three-winding transformer connected as stated above.

Thus we see that by judicious employment of three windings we can make better transformers than with only two windings. There are, of course, many applications of three-winding transformers where all three windings are separately used, each feeding separate circuits. These applications go far beyond the scope of this work, and it is necessary to mention only two.

Certain types of radio-frequency measurement bridges employ transformers, Fig. 20. The input transformer requires no special attention apart from screening between primary and secondary but the output transformer must be carefully studied. Two cases should be considered. When the turns ratio between AB and CB is not unity it would be desirable to have the zero partial leakage inductance in the lower impedance arm when this impedance is so low that inevitable wiring stray inductances might enter the problem. If the impedance level is not as low as this then it might be better to have the zero partial in the output arm, maintaining the turns ratio between the two bridge winding partials.

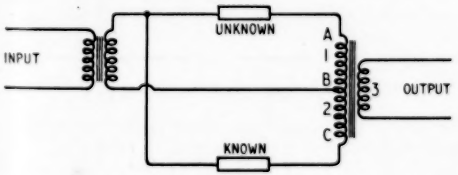



Fig. 20. *Illustrating the use of a transformer in a bridge.*

A case of a transformer required to connect three different circuits together which occurred in practice is shown in Fig. 21. A receiver had to be fed simultaneously from an 80-ohm dipole and a 150-ohm standard signal generator. The electromagnetic field to be measured was received on the dipole and produced a meter reading in the receiver. The dipole was then replaced by an 80-ohm resistor and the same receiver meter reading obtained with the aid of the standard signal generator. The mis-match into the receiver was intentionally arranged so that variations in receiver input impedance did not affect transformer loading on the feed circuits.

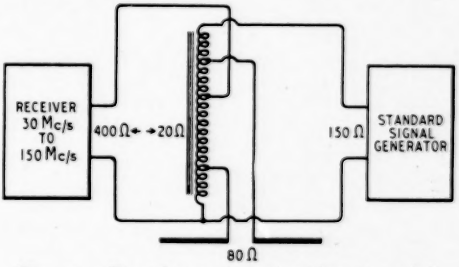



Fig. 21. *Transformer for connecting an aerial and a signal-generator to a receiver.*

The foregoing ideas on three-winding transformers may be extended to multi-winding transformers. However many windings are used, only one partial leakage inductance can be made zero. The rather tedious calculations may be summarized for the case of an $m$-winding (spiral) transformer of unity impedance ratios in which the windings are not interwound.

$$l_1 = \frac{4\pi n^2}{1000h} A_{12} \quad .. \quad .. \quad .. \quad (42.1)$$

$$l_2 = 0 \quad .. \quad .. \quad .. \quad .. \quad (42.2)$$

$$l_3 = \frac{4\pi n^2}{1000h} A_{23} \quad .. \quad .. \quad (42.3)$$

$$l_4 = \frac{4\pi n^2}{1000h} (A_{23} + A_{34}) \quad .. \quad (42.4)$$

---

$$l_m = \frac{4\pi^2 n}{1000h}(A_{23} + A_{34} + A_{45} + \ldots\ldots\ldots + A\,(m-1)m) \quad .. \quad (42.5)$$

## 7. Auto-transformers

This type of transformer is of very great utility, in fact most of the transformers designed by the authors come into this category. The reason for this is that, all other things being equal, an auto-transformer has lower leakage than its separate winding counterpart.

Consider Fig. 22 (*a*) and (*b*) which shows an auto-transformer and its equivalent separate winding transformer. Neglect all

shunt losses and re-draw in the form (*c*) and (*d*). The "not-ideal" transformer can be re-drawn as in (*e*). This assumes that the leakage referred to the $n$-turn side is the same in both the $n$ to $n+m$ and $n$ to $m$ transformers. As this means they both have the same coupling factors it is a reasonable assumption. By moving the leakage $l$ from primary to secondary we get (*f*).

The two ideal primaries in parallel may be drawn as one so Fig. 22 (*f*) becomes (*g*). Changing the leakage from secondary to primary we get (*h*). Thus the ratio of the leakage inductance of a separate-winding transformer to that of an auto-transformer is, comparing Fig. 22 (*h*) with (*d*).

$$\frac{\text{2-winding transformer leakage}}{\text{auto-transformer leakage}} = \left(1+\frac{n}{m}\right)^2 \quad .. \quad .. \quad .. \quad (43)$$

This advantage is nearly always worth having.

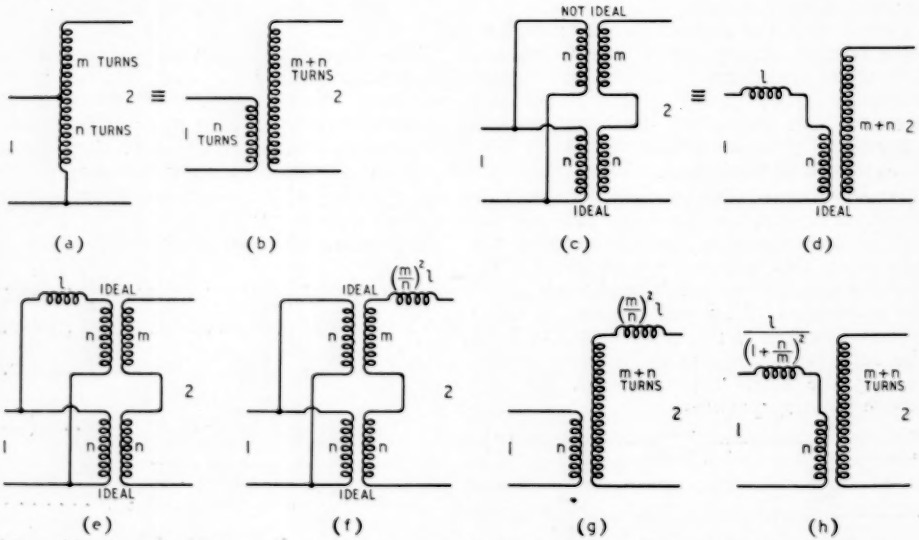



Fig. 22. *An auto-transformer (a) has the equivalent transformer (b). Neglecting shunt losses these can be drawn in the form (c) and (d). In the steps shown the final equivalent of (h) is reached.*

The above deals with an unbalanced-to-unbalanced case but a very frequent application is that of unbalanced-to-balanced transformers as in Fig. 23. In this case we have

$$\frac{\text{2-winding transformer leakage}}{\text{auto-transformer leakage}} = \frac{4\left(1+\frac{n}{m}\right)^2}{3\left(\frac{n}{m}\right)^2+2} \quad .. \quad .. \quad .. \quad .. \quad (44)$$

This equation does *not* hold when $\frac{n}{m}=0$

It may be seen from Fig. 23 (*a*) that this type of auto-transformer is a special case of a three-winding transformer. A quick method of determining the effective leakage inductance between the balanced windings and the unbalanced windings for purposes of deciding on the best positions for them is as follows: [see Fig. 23 (*a*) for winding numbers].

$$L = \frac{1+n/m}{2}\left(\frac{L_{13}-L_{12}}{n/m}+L_{23}\right) \quad .. \quad (45)$$

where $L$ is as shown in Fig. 23 (*b*) and $L_{13}$, $L_{12}$ and $L_{23}$ have the same significance as in Section 6 and equations (37).

## 8. Transformer Balance

When the transformer specification calls for one of the windings to be balanced and the other one not, it is desirable that the degree of balance should be at least 20 db and usually it is not necessary for it to exceed 40 db. There are two common types of balance required. In one case the transformer might be required to feed a balanced circuit and usually it is sufficient if [Fig. 24 (*a*)]

$$20\log_{10}\frac{V_A-V_B}{V_A+V_B} \leqslant -20 \text{ db} \quad .. \quad (46)$$

The second case is that in which a balanced circuit feeds an unbalanced circuit. In this case it is often necessary to have a somewhat better balance. Consider Fig. 24 (*b*) wherein

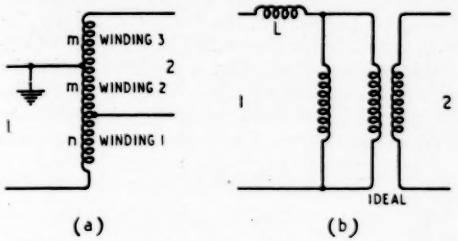



Fig. 23. *Unbalanced-to-balanced auto-transformer (a) with its equivalent (b).*

a balanced aerial feeds a coaxial cable. Normally there will be unbalanced or push-push voltages represented by $e$. If the transformer balance were perfect there would be no voltage across the coaxial cable due to $e$. The degree of balance may well be measured in the manner shown in Fig. 24 (*c*). A voltage $V$ from a standard signal generator represents the aerial push-push voltage. $V_1$ is the undesired unbalanced voltage due to $V$. Normally the transformer may be considered satisfactory if

$$10 \log_{10} \frac{\frac{V_1^2}{R_1}}{\frac{V^2}{R}} = 20 \log_{10} \frac{V_1}{V} + 10 \log_{10} \frac{R}{R_1} \leqslant -25 \text{ db} \quad .. \quad .. \quad .. \quad (47)$$

Unbalance is due to asymmetry of winding and of shunt capacitances between the two halves of the balanced winding, and to electrostatic coupling between the balanced and unbalanced windings. Asymmetry is obvious and must be avoided as far as possible. The electrostatic coupling capacitances are shown in Fig. 24 (*d*).

$C_1$ and $C_2$ can be rendered harmless by the interposition of a copper screen between the unbalanced and balanced windings. Care must then be taken that they are either small or equal, otherwise they will cause flux due to currents caused by $e$. The screen may be dispensed with if the leakage inductance between the two halves of the balanced winding be made small enough. The voltages at $a$ and $b$ due to push-push currents are the same as at o, namely zero. If $L_2$ and $L_3$ are small enough the voltages at A and B will likewise be zero. An entirely adequate balance for most normal purposes can be obtained without the use of screens by using concentric spiral windings of wide strip. Usually the leakage between the two halves is so small that the presence of $C_1$ and $C_2$ is not troublesome. If the above leakage is not so small, $C_1$ and $C_2$ can be made practically zero by careful positioning of the windings.

## 9. Examples.

As an example of the application of some of the foregoing information we shall design a transformer to fulfil the following specification :

Impedance Ratio : 100 ohms unbalanced to 100 ohms balanced and centre-tapped.

Frequency Range : 50 kc/s to 75 Mc/s.

Insertion Loss : to be not greater than 2 db.

Power Handling Capacity : to be not greater than one watt.

Primary-Secondary Insulation : not insulated.

Phase Response : unimportant.

Use : balanced cable to unbalanced cable, balance cable to receiver input, etc.

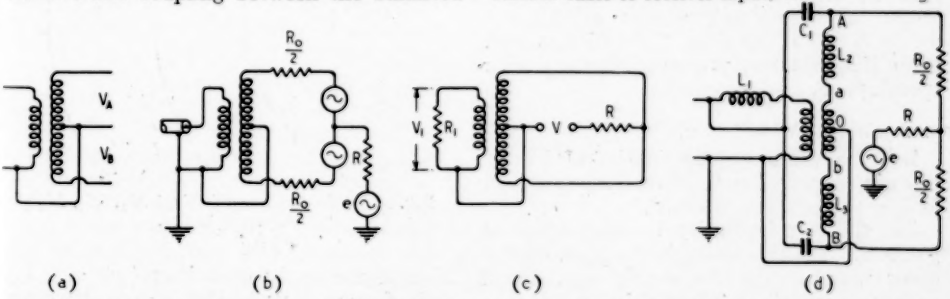



Fig. 24. *A balanced transformer has the form (a). If used to feed an unbalanced cable (b), push-push voltages e will produce a current in the cable unless the balance is perfect. This is shown at (c) and the effect of capacitance at (d).*

9.1. *Design.*

The primary-secondary insulation requirement allows the use of an auto-transformer of the type shown in Fig. 23(*a*). We shall adopt a toroidal core, and inspection of Fig. 5 shows that Mumetal permits of the lowest number of turns for a 2-db insertion loss at 50 kc/s. From Fig. 5 we see that fourteen turns for each 100-ohm winding will suffice. As we shall make it an auto-transformer the total number of turns need be twenty-one only, as one seven-turn winding will be in both the primary and secondary circuits. The advantage we gain by the use of an auto-transformer is, from

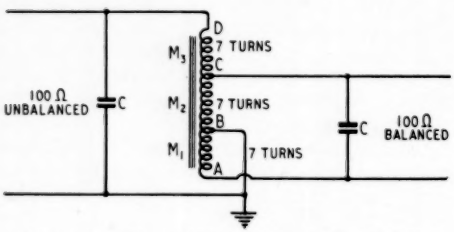



Fig. 25. *The auto-transformer of which the design is given as an example.*

equation (44), 3.2 times less leakage. We shall want the lowest possible leakage between the two halves of the balanced secondary so let us use concentric spiral windings of flat strip with the two half-secondary windings inside, and the half-primary winding outside. As we want minimum balanced-winding leakage and the electrostatic coupling capacitances will be practically zero because of the concentric spiral type of winding we shall use the widest possible strip and check that the shunt capacitances will not be too great. If this check proves satisfactory we shall then calculate the leakages and again if satisfactory proceed with the design specification.

9.2. *Shunt Capacitances.*

Let the winding strip consist of $\frac{1}{2}$ in $\times$ 0.001 in copper. Let the interturn insulation be $\frac{3}{4}$ in $\times$ 0.001 in rice tissue.

Let the core insulation be four layers of $\frac{1}{4}$ in $\times$ 0.001 in rice tissue half-lapped on the outer periphery of the toroid and doped with polystyrene. The thickness of core insulation therefore becomes 0.001 in $\times$ 4 $\times$ 2 = 0.008 in on the outside of the toroid and 0.016 in on the inside; say an average thickness of 0.012 in. The core cross section is $\frac{1}{4}$ in $\times$ $\frac{1}{4}$ in so that the periphery of the cross section of the insulated core becomes 4 (0.25 + 2 $\times$ 0.012) = 1.096 in. If we assume this core cross section to be a circle of equal periphery, the radius will be $r_1$ = 0.175 in. This is the radius of the start of the balanced winding, Fig. 25. The radius of the finish of this winding, that is the radius of the point C will be 0.175 + 14 $\times$ 2 $\times$ 0.001 = 0.203 in. The average radius will be $\frac{0.175 + 0.203}{2} = 0.189$ in. The interturn capacitance, called $C_2$ in Section 4.2 is evidently

$$C_2 = \frac{K \times \text{Area}}{4\pi \times \text{Distance}} = \frac{2.3 \times 2\pi \times 0.189 \times 5}{4\pi \times 0.001} \times 2.54 \times 1.1 \text{ pF}$$

$$C_2 = 304 \text{ pF}$$

From equation (29)

$$C = \frac{14 - 1}{14^2} \times 304 \text{ pF}$$

$$C = 20 \text{ pF}$$

The capacitance between B and D, Fig. 25, will be slightly higher than this because the average winding radius will be greater, but such a small difference as this is not worth taking into account in view of the poor precision of shunt capacitance calculations.

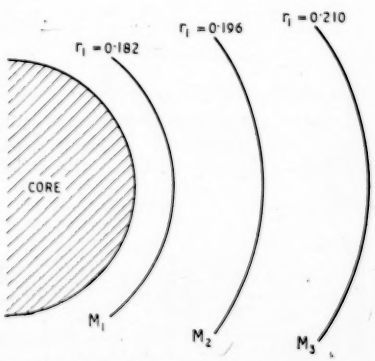



Fig. 26. *The core and windings $M_1$, $M_2$, $M_3$ of the transformer.*

Now if we apply equations (32) and (33) with $f_2$ = 75 Mc/s and $R_0$ = 100 Ω we find that $C$ = 21 pF so that we shall just succeed in our design if we adopt the low-pass filter technique in Section 5. The maximum

permissible leakage inductance is from equation (33) $L = 0.21\,\mu H$. This must be calculated from equation (45) when we are in a position to know the short circuit inductances $L_{12}$, $L_{23}$ and $L_{13}$.

9.3. *Leakage Inductance.*

We shall consider each of the three windings $M_1$, $M_2$ and $M_3$ as if they had no radial thickness and assume them to be concentrated at a distance from the centre of the core cross section equal to the mean radius of each winding, Fig. 26.

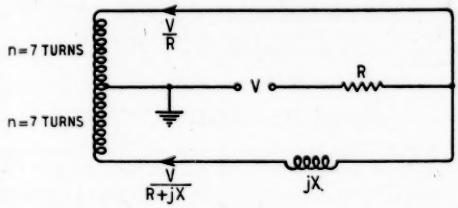



Fig. 27. *The leakage inductance of the balanced winding appears on one side only.*

Now apply equations (39) and (37): $h = \frac{1}{2}$ in ; $A_{12} = 2\pi \times 0.189 \times 0.014 = 0.0166$ in$^2$ ; $A_{23} = 2\pi \times 0.203 \times 0.014 = 0.0178$ in$^2$ ; $A_{13} = A_{12} + A_{23} = 0.0344$ in$^2$ ;

$$L_{12} = \frac{4\pi \times 7^2 \times 0.0166}{1000 \times \frac{1}{2}} \times 2.54 = 0.052\,\mu H ;$$

$$L_{13} = \frac{4\pi \times 7^2 \times 0.0344}{1000 \times \frac{1}{2}} \times 2.54 = 0.107\,\mu H ;$$

$$L_{23} = \frac{4\pi \times 7^2 \times 0.0178}{1000 \times \frac{1}{2}} \times 2.54 = 0.0557\,\mu H.$$

Now apply equation (45) wherein $n = m$

$$L = 2L_{23} = 0.11\,\mu H$$

which is well below the maximum permissible, $0.21\,\mu H$. It might be advisable in view of the closeness of the capacitance to the limiting value of 21 pF and the considerable tolerance on the leakage, to investigate a re-design using thinner strip, say $\frac{1}{4}$ in wide instead of $\frac{1}{2}$ in. This is not worth trying, however, until a model has been made up based on the present design, because one must not rely on the capacitance and leakage inductance formulae to give an actual working design but only as an indication that a design is feasible.

9.4. *Balance.*

As previously stated the electrostatic coupling between windings is negligible. Lack of balance will be due to actual asymmetry in the balanced winding. Consider the three partial inductances obtained by regarding the transformer as a three-winding device. Equations (40) give

$$l_1 = L_{12} = 0.052\,\mu H$$
$$l_2 = 0 \text{ very nearly}$$
$$l_3 = L_{23} = 0.056\,\mu H.$$

From this it may be seen that the leakage between the two halves of the balanced winding is entirely on one side of the centre-tap and equal in value to $0.052\,\mu H$. Thus the circulating current in the balanced winding due to a push-push voltage $V$, Fig. 27, is

$$\frac{V}{R} \frac{1}{\sqrt{\left(\frac{R}{X}\right)^2 + 1}}$$

This current will appear on the unbalanced winding and will cause a voltage to appear

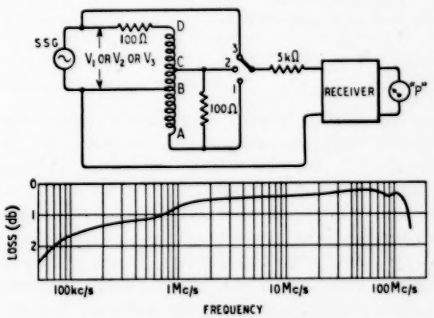



Fig. 28. *Frequency response of the auto-transformer described. The receiver output P was kept constant and the signal generator output varied ; $V_1$, $V_2$, and $V_3$ were found by connecting the receiver input to 1, 2, and 3 respectively.*

across the terminating resistor $R_1$ (100 Ω in our case) equal to $V \frac{R_1}{R} \frac{1}{\sqrt{\left(\frac{R}{X}\right)^2 + 1}}$.

The unbalance ratio as described by equation (47) becomes

$$20 \log_{10} \frac{R_1}{R\sqrt{\left(\frac{R}{X}\right)^2 + 1}} + 10 \log_{10} \frac{R}{R_1}$$

If we let $R_1 = R = 100\ \Omega$ and $X$ be the reactance of 0.052 $\mu$H at 75 Mc/s the above ratio becomes $-12\frac{1}{2}$ db. This is rather poor discrimination against push-push voltage but not so bad as to cancel the making of a first model. A test model transformer was made

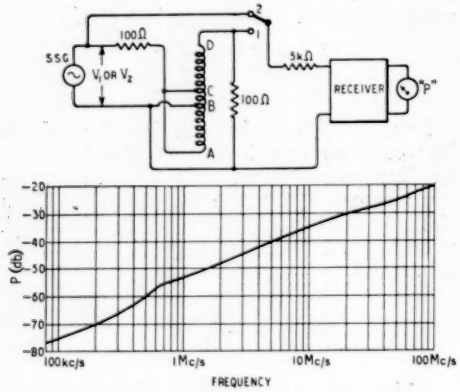



Fig. 29. *Unbalanced voltage ratio of the transformer* $= P_{db} = 20\ log_{10}\ (V_2/V_1)$. *The receiver output P was kept constant and the signal generator output varied; $V_1$ and $V_2$ were found by connecting the receiver input to 1 and 2 respectively.*

on the above lines. The shunt capacitance turned out to be 10 pF instead of the calculated 20 pF. The leakage between balanced and unbalanced windings was actually 0.12 $\mu$H instead of the calculated 0.11 $\mu$H. The unbalance ratio was $-22$ db at 75 Mc/s instead of the calculated $-12\frac{1}{2}$ db. The combination of the actual measured shunt capacitance and leakage inductance is such that the transformer cut-off becomes 150 Mc/s instead of 75 Mc/s.

Fig. 28 shows the insertion loss of the transformer while Fig. 29 shows the unbalance ratio.

## 10. Acknowledgments

The basic principles underlying the foregoing work are well known but their application to transformer design was first shown to the authors by C. G. Mayo to whom much is owed, not only as regards technique but also courage to apply the basic principles of audio-frequency transformer design to very-high frequencies.

The authors wish to thank Miss M. Richards for making the many tedious measurements necessary for the compilation of Figs. 5, 6, 7, and 8. Much assistance was given by D. P. Thurnell, particularly as regards the making and testing of the example transformer. The authors are very grateful to H. L. Kirke for the continual interest and encouragement which they have received. Their thanks are also due to The British Broadcasting Corporation for kind permission to publish this paper.

### REFERENCES

[1] "Magnetic Circuits and Transformers," by Members of the Staff of the Dept. of Electrical Engineering of Massachusetts Institute of Technology.

[2] "Equivalent Capacitances of Transformer Windings," by W. T. Duerdoth, *Wireless Engineer*, June 1946.

[3] "Mathematics Applied to Electrical Engineering," by A. G. Warren, for formulae relating to interwinding capacitances, paragraph 8.7.

## HONOURS AND AWARDS

In the Birthday Honours, the following have been appointed Commanders of the British Empire :—

**O. F. Brown.** Headquarters Intelligence Division of the Department of Scientific and Industrial Research.

**H. L. Kirke.** Head of the Research Department of the British Broadcasting Corporation.

The award of the Medal for Merit was recently made by the Government of the United States to a number of British Scientists for "exceptionally meritorious conduct in the performance of outstanding services to the United Nations." Listed below are some of those awarded the medal, with extracts from the citations :—

**Sir Edward Appleton.** "Organized and supported scientific workers, prosecuting many of the important scientific projects which resulted in Allied supremacy. He was outstanding in the effectiveness of his work for co-operation between British and American scientists, and was responsible for important phases of the early research leading to the development in England and the United States of modern radar."

**Dr. Robert Cockburn.** "Was instrumental in devising new electronic techniques to counter the enemy's use of fire control methods, navigational aids and bombing devices."

**Albert Percival Rowe.** "Exercised a dominant role in the development of radar and countermeasures in the United Kingdom. In the development and employment of electronics in warfare he made important contributions."

**Sir Robert Watson Watt.** "His years of work . . . largely formed the basis of the extraordinary early success of British radar. . . . The success attained in the military employment of these new scientific devices and methods inspired and stimulated the great radar programme in the United States. . . . In the development and effective combat employment of superior types of radar, radar countermeasures and communications equipment his services were of enormous value."

# CORRESPONDENCE

*Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.*

## Valve Equivalent Circuit

*To The Editor, "Wireless Engineer,"*

SIR,—I should like to comment on the question of the relative validity of the constant-voltage and constant-current generator "equivalent" circuits of the thermionic valve, raised by Dr. Salzberg in the April issue of *Wireless Engineer*.

I believe the origin of the alternative representations to be two-fold. First, analysis of the constant-voltage generator circuit (Dr. Salzberg's Fig. 1) yields

$$|i_p| = \frac{\mu}{r_p + R}|e_g| \quad .. \quad .. \quad .. \quad .. \quad (1)$$

from which

$$|e_o| = R|i_p|$$

$$= R \cdot \frac{g_m r_p}{r_p + R}|e_g|, \text{ using } \mu = g_m r_p$$

$$= \frac{g_m}{y_p + Y}|e_g| \quad .. \quad .. \quad .. \quad .. \quad (2)$$

The correspondence between expression (1) above and Fig. 1, suggests the possibility of setting up a configuration equivalent to expression (2); this is the origin of Fig. 2. This development has been given in one form or another in several standard texts; e.g., F. E. Terman, *Radio Engineering*, p. 173. While (2) has been derived from (1) the derivation could with equal validity be reversed. It should be noted that the two expressions differ by a constant factor $R$ and cannot therefore have exact equivalence. A further point is that $r_p$, $y_p$ are regarded as passive elements in these equivalent circuits; the significance of this point will appear later.

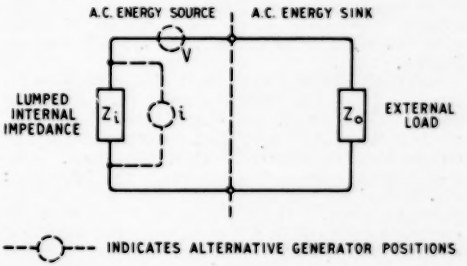
It is of interest to observe that the two expressions and their respective equivalent configurations are related by the principle of duality. This fact arises naturally since (1) is concerned with current distribution while (2) expresses the distribution of voltage; moreover it explains the agreement between expressions (2) and (4) of Dr. Salzberg's letter in the special case $r_p = R$. Duality does not, of course, necessarily imply even partial equivalence.

The second origin of the two "equivalent" circuits is provided by the Helmholtz-Thévenin Theorem on which basis a source of e.m.f. may be dissociated from its impedances and the latter lumped into a single equivalent two-terminal impedance. Two possibilities arise here since the generator may be connected either in series or in parallel with its own equivalent impedance according as it (the pure generator) is imagined to have zero or infinite impedance respectively.

Turning now to the discrepancy noted by Dr. Salzberg, it cannot be expected that the total power absorbed in the two circuits will be identical if only the one variable, the input voltage, is made the same for each circuit. Equal power distribution requires equal input powers; i.e., $(\mu e_g)^2/(r_p + R)$ for Fig. 1 must be numerically equal to $(g_m e_g)^2/(y_p + Y)$ for Fig. 2. Using $g_m/y_p = \mu$, this equality can only be satisfied when $I_p = R$, or $y_p = Y$.

With regard to the physical processes occurring in an actual valve circuit, so lucidly treated in your editorial of the same issue, the following appear to be the significant facts from the point of view of setting up an equivalent linear configuration. First, assuming linear negative-grid operation, the signal source contributes no power to the a.c. output circuit. Secondly the mean power supplied to the output circuit by the d.c. source is independent of signal amplitude. Finally, the energy released by the application of the signal arises, and is dissipated, within the same (the output circuit) loop. These features allow the d.c source to be omitted from the equivalent circuit.

Coming now to the correct method of arriving at the equivalent circuit it will be convenient to obtain first the correct impedance conditions, on the basis of the Helmholtz-Thévenin Theorem. The result is shown in the figure. The separation of the




active generating element from its inherent impedance(s), for the purpose of facilitating analysis, allows the alternative methods of connection indicated, provided the series generator has zero impedance and the shunt generator has infinite impedance. Theoretically it would appear equally valid to inject either e.m.f. or current by either method, provided the transfer constant relating the magnitude and phase of the injected quantity to the grid voltage be appropriately chosen. Of the two methods of injection the series method presents no difficulties; the shunt method, however, is open to ambiguous interpretation since the generated current may be imagined to flow through the hypothetically separate infinite impedance generator or through the parallel finite impedance path representing the inherent impedance of the source being represented. I believe that the first possibility is conventionally accepted, presumably following Wigge's Dual Theorem, with the result noted by Dr Salzberg, and would suggest that the second interpretation is correct since the finite parallel impedance must logically dissipate the power lost within the actual generator and must therefore pass the entire output current. In other

words, from the energy view point $r_p$ is an active element in the equivalent circuits.

It is not suggested that Wigge's theorem is incorrect but that it is not applicable to the accompanying figure since $z_i$ represents an *active* impedance.

In conclusion, I feel that teachers of radio theory, to whom problems of the kind under discussion are of the utmost importance, would be extremely grateful if you would leave your columns open on this question until definite clarification has been obtained.

A. W. KEEN

Harrow, Middx.

*To the Editor "Wireless Engineer."*

Sir,—In my previous letter[1] I pointed out that the two well-known first-order a.c. equivalent circuit representations for the vacuum tube yield identical results for the voltage across, current through, and power developed in the external impedance, and dissimilar results for the voltage across, current through, and power developed in the internal resistance. I stated that the result for the power developed in the internal resistance, as computed on the basis of the constant-voltage generator equivalent circuit was the correct one, and furthermore that this circuit is more fundamental than the alternative constant-current generator circuit. The objects of my present letter are to discuss the matter further, and to present a revised conclusion. Re-consideration of the subject was prompted by an exchange of views with Dr. H. Krutter of the N.R.L. Field Station, Boston.

Although the two equivalent circuit representations do indeed provide dissimilar results for the power developed in the internal resistance, one result is no more correct than the other, nor is one circuit more fundamental than the other. One reason for this revised conclusion is the following: Eq. (10) of the previous letter expresses $i_p$ in two different ways. The first is an expression for current through $Z$ in a constant-current generator circuit which contains the shunt elements $r_p$ and $Z$. The second is an expression for current in a constant-voltage generator circuit which contains the series elements $r_p$ and $Z$. The two circuit representations are arrived at simultaneously, and one circuit need not be regarded as having been derived from the other on the basis of equivalence with respect to voltage across and current through the external impedance. Consequently, it cannot be concluded that one circuit is more fundamental than the other, nor that the result yielded by one for the power developed in the internal resistance is more correct than the dissimilar result yielded by the other.

At first thought, it may appear that this curious lack of uniqueness between the two equivalent-circuit representations may be resolved by a consideration of the tube plate dissipation, a matter which was not discussed in the previous note. In general the tube plate dissipation is equal to the difference between the power supplied by the plate battery and that developed in the external impedance. Since the two circuit representations provide identical expressions for the power developed in the external impedance, as stated previously[1], both lead to the same value of plate dissipation.

It has been pointed out[2] that the power developed in the external impedance is equal to the difference between the total power supplied by the voltage generator of the constant-voltage generator circuit and that developed in the internal resistance; and furthermore, that these two terms may be interpreted as contributing to plate cooling and to plate heating, respectively. Equally well, it may be pointed out that the power developed in the external impedance is equal to the difference between the total power supplied by the current generator of the constant-current generator circuit and the corresponding power developed in the internal resistance, with a similar interpretation. Both the total generated power and the power developed in the internal resistance, as computed on the basis of the two circuit representations, are dissimilar. Nevertheless, the results yielded by one circuit are no more correct than those yielded by the other. As far as all external measurements are concerned, including plate dissipation, both circuit representations provide identical results.

The situation for the vacuum tube is quite unlike that which exists in the following analogous case: Suppose it were required to determine whether a given two-terminal source is a constant-voltage or constant-current generator. Consideration will reveal that no measurement of voltage across, current through, or power developed in an external impedance connected to the two terminals could be used to decide the matter. However if a measurement could be made by a calorimeter experiment, say, of the power developed within the internal resistance of the device, with the external terminals either open-circuited or short-circuited, the answer would be determined immediately.

Washington, D.C., U.S.A. BERNARD SALZBERG. Naval Research Laboratory.

[1] Bernard Salzberg, "On Use of Equivalent Circuits to Represent the Vacuum Tube," *Wireless Engineer*, Vol. xxiv, No. 283, p. 124, April 1947.
(See also April Editorial).

[2] Harry Stockman, "The Validity of the Equivalent Plate-Circuit Theorem for Power Calculations," *Proc. Inst. Radio Engrs*, June 1944, Vol. 32, No. 6, p. 373.

## Permeability of Dust Cores

*To the Editor, "Wireless Engineer."*

SIR,—At the risk of boring your readers by perhaps over-working the analysis of permeability of magnetic dust cores, I am rising to the bait cast by yourself and several letter-to-the-editor writers.[1] It appears that the chief subject of debate is the failure of nature to conform to simple mathematical analysis. This is always disappointing to physicists, or to engineers, who, for example, prefer Boyle's Law to more accurate, if more obscure, statements of the behaviour of gases.

It is noteworthy that several correspondents are disturbed or see anomaly in the failure of a ferromagnetic body to conform to cubical or spherical patterns which can be treated by straightforward mathematical means. Ferromagnetism is a veritable morass for such efforts, and as often as not it finally compels its votaries to resort to the empirical approach, if anything tangible is to be achieved in their lifetimes.

The permeability of a dust core depends indeed upon the permeability of the material of each magnetic particle, and upon the effective airgap in the magnetic circuit. But neither of these dimen-

sions is directly accessible to our knowledge in a practical core.

The initial permeability of mechanically hard-worked molybdenum - permalloy is about 50, while that of well-annealed, stress-free material is several thousand. The annealing treatment of molybdenum-permalloy dust cores is not sufficient to develop the highest intrinsic permeability, nor to remove all the mechanical stresses which depress permeability. We thus infer that the intrinsic permeability of dust-core particles lies well above 50, but we cannot infer from our knowledge of the mechanical and heat treatments how much higher it is.

The particle size and shape vary widely, depending upon the grain structure in the permalloy as originally cast, and upon the method and intensity of the pulverization process. Thus magnetic particles of various shapes and sizes are present, together with insulating material, which, it is hoped, is uniformly spread on the surfaces of the magnetic particles. The force used in compressing the core is sufficient to cause metallic flow in all obstinate particles which attempt to bridge gaps, or which resist nesting together with other particles. The force may be enough in spots to drive metal protuberances right through the insulating layers, and into direct contact with each other (to make magnetic bridges). And after that the annealing treatment serves to sinter neighbouring metallic particles together wherever possible. This introduces other stresses and discontinuities in the sensitive magnetic alloy, and it yields sinuous flux paths, with miscellaneous air-gaps and flux concentrations.

It is evident from the above that any analysis of magnetic behaviour will entail such great complication as to be unprofitable. The empirical equation $\mu = \mu_i{}^p$ of the Legg-Given paper was propounded as useful and simple. It gave the permeability of molybdenum-permalloy dust cores with tolerable accuracy for a wide range of conditions of core pressure and dilution, ranging from $\mu = 3$ up to $\mu = 125$. It requires only two parameters, the "intrinsic permeability" $\mu_i$, and the metallic packing factor $p$. Such a simple and reliable equation is worthy of respect and use despite any difficulty in tracing its theoretical antecedents.

The extrapolation to $\mu_i = 220$ yields a value of intrinsic permeability which is at least reasonable, in view of the minimum and maximum limits on permeability already discussed. The estimate $\mu_i = 1{,}000$ made by Bardell on the basis of a core pressed with no insulation is not reliable, since the amount of sintering, and resultant easy magnetic paths on the one hand, and residual internal stresses on the the other hand, complicate the resultant permeability. This compels resort again to the relationship which has been found to apply for cores as insulated and as annealed, without any invalidating assumptions.

Derivation of a formula for core permeability from assumptions closely related to the complex physical facts remains as a problem for those who are dissatisfied without such derivations. This may be expedited by the time-honoured procedure of "working backward" from a known correct answer, as provided by our equation.

V. E. LEGG.

New York. Bell Telephone Laboratories.

[1] *Wireless Engineer*, February 1947, pp. 33 and 63; November 1946, Vol. XXIV, pp. 291 and 313

## Degrees for Ex-Servicemen

*To the Editor, "Wireless Engineer."*

SIR,—With reference to my letter in the April issue on Science and Engineering Degrees for Ex-Servicemen and others, I have received a total to date of ninety-three completed questionnaires. Of these, fifty-one are from the London area, including Dartford and South-West Essex, and ten from Stoke-on-Trent. The numbers in each of the other localities are unfortunately too small to make it possible to ask the Ministry of Education to take action, but it is hoped that in London and Stoke it may be possible to bring about the desired results.

I am communicating with the Ministry of Education, to whom the questionnaires and an analysis of the results are being sent.

I should like very much to thank those who have completed questionnaires and to request those who have received them but who have not yet returned them to do so immediately, especially if they wish to take courses in the London area or in Stoke.

Langley Park, Nr. Slough, Bucks. O. S. PUCKLE, R.F. Equipment, Ltd.

## 1947 RADIO CONVENTION

The British Institution of Radio Engineers held a convention at Bournemouth from 19th to 23rd May. The following papers were read :—

Transmitting Valves for Communication on Short Wavelengths, by W. H. Aldous.
Television Receiving Aerials, by D. A. Bell.
The Exploitation of Micro-Waves for Trunk Waveguide Multi-Channel Communications, by Prof. H. M. Barlow.
A One Kilowatt V.H.F. Frequency Modulated Transmitter, by J. B. Lovell Foot.
International Telegraph Networks, by Dr. E. V. D. Glazier.
Ultra High Frequency Modulation on Wave Guide, by Drs. H. Gutton and J. A. Ortusi.
Automatic Audio Frequency Response Curve Tracer, by G. L. Hamburger.
Functional Requirements for Radio Aids to Civil Aviation, by Capt. V. A. M. Hunt.
The Klystron as Amplifier at Centimetric Wavelengths, by R. Kompfner.
The Problems of Radio Communication with Moving Trains, by G. H. Leversedge.
The Allouis (France) Short-Wave Broadcasting Centre, by Dr. M. Matricon.
A Direct Reading Frequency Measuring Set, by F. C. F. Phillips.
Radio Navigational Aids, by W. J. O'Brien.
International Automatic Telegraph Networks, by Commander J. D. M. Robinson, R.N.
The Broadcast Antenna, by Dr. H. Paul Williams.

## PROFESSOR G. W. O. HOWE

On 25th June, Prof. G. W. O. Howe received the honorary degree of Doctor of Laws from the University of Glasgow. Some months ago he retired from the Chair of Electrical Engineering of the University.

# WIRELESS PATENTS

## A Summary of Recently Accepted Specifications

*The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.*

### DIRECTIONAL AND NAVIGATIONAL SYSTEMS

581 057.—Filtering arrangement for increasing the signal-to-noise ratio, when receiving the reflected echo-signals in radiolocation, or in pulsed communication systems.

*Standard Telephones and Cables Ltd. (assignees of E. Labin and R. B. Hoffman). Convention date (U.S.A.) 9th September, 1943.*

581 120.—System for locating the instantaneous position of a moving target, and predicting its course, by the use of a continuous exploring wave.

*Standard Telephones and Cables Ltd., B. B. Jacobsen and B. Secker. Application dates 12th December 1939, and 7th March, 1940.*

581 126.—Dynamometer indicator for blind-approach systems of the overlapping-beam type.

*Standard Telephones and Cables Ltd. and C. W. Earp. Application date 26th December, 1941.*

581 165.—Radiolocation system in which (*a*) exploring pulses are generated by a super-regenerative oscillator, and (*b*) the reflected pulses are received during the quiescent intervals and also determine the quenching frequency.

*Marconi's W.T. Co. Ltd., J. M. Furnival and N. M. Rust. Application dates 1st April and 1st September, 1941.*

581 160.—Range-finding system in which the target is explored (*a*) by a frequency-modulated and (*b*) by an unmodulated wave, the reflected echoes being fed to separate receivers.

*Nash and Thompson Ltd., A. G. Frazer-Nash and A. Whitaker. Application date 8th November, 1939.*

581 161.—Balanced impedance bridge for measuring small capacitances, particularly in an arrangement for indicating the altitude of an aeroplane in terms of its capacitance to ground.

*A. D. Blumlein. Application dates 10th January, 11th March and 17th June, 1940.*

581 166.—Determining the speed of a radio-pulsed target by passing part of the echo-signal through a delay circuit, and then mixing that signal with an undelayed signal.

*J. Forman and Pye Ltd. Application date 17th April, 1941.*

581 167.—Quarter-wave transmission-line bridge for balancing stray capacitances, for instance in radiolocation equipment.

*E. C. Cork and A. D. Blumlein. Application date 5th May, 1941.*

### TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

581 018.—Television system in which a two-stage limiter "clips" the video-signal components in order to provide a clear-cut synchronizing-signal.

*Haseltine Corporation (assignees of J. A. Rado). Convention date (U.S.A.) 14th May, 1943.*

### TRANSMITTING CIRCUITS AND APPARATUS

*(See also under Television)*

580 378.—Waveguide made of synthetic resin, the inner surface being first coated with metal-foil and then electroplated.

*C. E. Fenwick and C. S. Wright. Application date 28th March, 1944.*

580 528.—Heterodyne arrangement for ensuring stability in intercommunication systems operating over a wide range of predetermined wavelengths.

*Soc. Française Radio-Electrique. Convention dates (France) 26th January and 3rd August, 1940.*

### SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

581 005.—Band-pass filter-circuit for converting time-modulated signal pulses into a phase- or frequency-modulated wave.

*Standard Telephones and Cables Ltd. and C. T. Scully. Application date 19th May, 1943.*

581 218.—Diversity-reception, or other multiple-path system, in which the harmonics produced by a limiter valve in one path control the gain-characteristic of another path.

*Western Electric Co. Inc. Convention date (U.S.A.) 1st July, 1943.*

### CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

581 481.—Velocity-modulation device comprising an electron-discharge tube coupled transversely to the longitudinal axis of a wave-guide.

*Standard Telephones and Cables Ltd. and S. G. Tomlin. Application date 29th August 1941.*

### SUBSIDIARY APPARATUS AND MATERIALS

580 844.—Generating square-wave triggering-voltages by a circuit comprising a master-oscillator and two slave-oscillators.

*M. E. Haine and Metropolitan-Vickers Electrical Co. Ltd. Application date 8th January, 1943.*

580 891.—Time-base circuit for a cathode-ray indicator, with means for "expanding" a selected portion of the sweep, so as to enlarge the image or trace under observation.

*J. M. Debski. Application date 28th June 1944.*

581 613.—Multivibrator of the pentode type with means for "gating" a selected signal-pulse, and for preventing fortuitous triggering.

*Standard Telephones and Cables Ltd. and C. W. Earp. Application date 15th September, 1944.*